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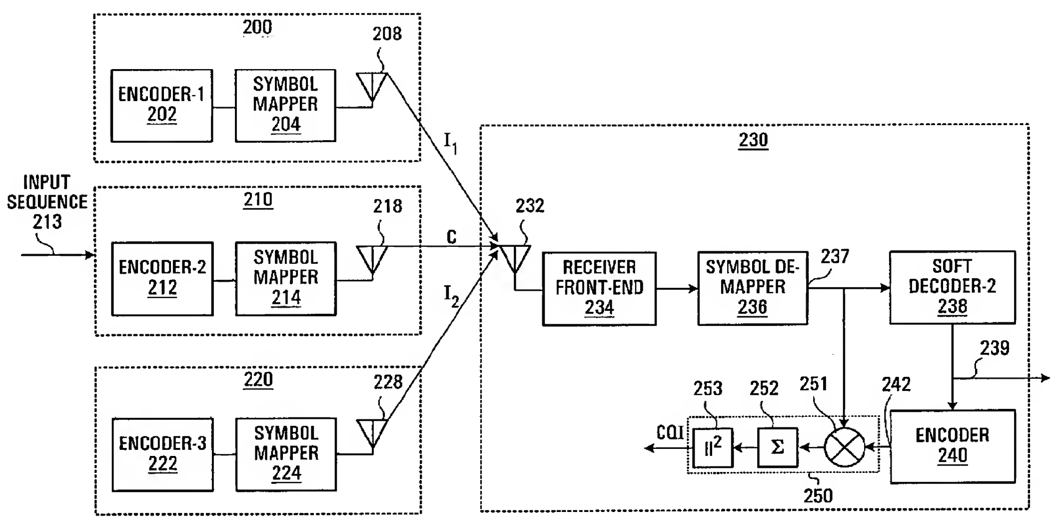
(54) Title: METHOD AND APPARATUS FOR CHANNEL QUALITY MEASUREMENTS



(57) Abstract: A method and apparatus are provided for combining pilot symbols and Transmit Parameter Signalling (TPS) channels within an OFDM frame. The method uses Differential Space-Time Block Coding to encode a fast signalling message at an OFDM transmitter. At an OFDM receiver, the encoded fast signalling message can be decoded using differential feedback to recover information about the channel responses that would normally be carried by pilot symbols. In wireless data transmission employing adaptive modulation and coding, an instantaneous channel quality measurement, independent of the origin of interference for example, neighboring-cell interference, white thermal noise, or residual Doppler shift is provided. Using the correlation between a signal which has been symbol de-mapped, and one which has also been soft decoded and re-encoded, a channel quality indicator is produced. Another embodiment uses TPS data as pilot symbols by decoding TPS and then re-encoding.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

METHOD AND APPARATUS FOR CHANNEL QUALITY MEASUREMENTS

FIELD OF THE INVENTION

The invention relates to wireless data transmission, and more particularly to channel quality measurement in respect of such data transmission.

BACKGROUND OF THE INVENTION

Adaptive modulation and coding is a key enabling concept and technology for high-speed wireless data transmission. A wireless channel is typically a random fading channel. Adaptive coding and modulation is a commonly employed solution for transmitting data over such an unknown channel. Conventional design methodology provides a large fade margin in the transmit signal power to combat deep fades which may occur. Such fade margins are typically at least 6dB, which represents a 200-300% throughput loss. The aim of adaptive coding and modulation is to fully utilize the channel capacity and to minimize the need to use such a fade margin by dynamically selecting the best coding and modulation configuration on-the-fly. This requires the transmitter to have accurate information about the instantaneous channel quality. Such instantaneous channel quality information is extracted at the receiver and fed back to the transmitter. The conventional approach is to measure the channel (signal) to interference power ratio (CIR) at the receiver front-end. Based on the instantaneous CIR and a targeted performance, the transmitter determines and applies the appropriate coding rate and modulation. In general, due to a complex propagation environment, a fast and accurate measurement of the CIR is a very difficult task.

Conventional channel quality measurements can be classified into two categories: (1) pilot based channel quality

measurements and (2) decision feedback based channel quality measurements. These methods use the correlation of known sequences, typically Pseudo-Noise (PN) codes, with both the desired signal and the interference. For a slowly varying channel with a sufficient measurement time, the conventional methods can provide an accurate CIR measurement.

Referring to figure 1, a conventional pilot based CIR estimation scheme will now be described. In the context of MIMO-OFDM (Multiple Input Multiple Output - Orthogonal Frequency Division Multiplexing), the conventional channel quality measurement uses a pilot header containing two identical known OFDM symbols upon which to base an indication of the current channel quality. Figure 1 shows a first, second and third base transceiver station (BTS) 100, 110, and 120 transmitting their respective signals, and a mobile station 130 receiving these signals. Mobile station 130 is configured to receive, demodulate and decode a signal transmitted by the second base transceiver station 110. The signals transmitted by the first base transceiver station 100 and the third base transceiver station 120 are received as interference by the mobile station 130. A channel associated with the signal having received signal power C transmitted by base transceiver station 2 (BTS_2) 110 is the channel whose quality is to be measured. Suppose that we have N PN codes, and that the length of each PN code is N chips, we then have:

$$\begin{array}{ll} PN_i \cdot PN_j \approx 0 & i \neq j \\ PN_i \cdot PN_i = N & 1 \leq i \leq N. \end{array}$$

This important relation that the PN codes form a near orthogonal set allows for the extraction of specific channels using the Pilot channel PN codes. In figure 1 only three BTSs are shown, and hence there are only three PN codes. The second BTS 110 encodes a signal whose associated channel quality is to

be measured, at ENCODER-2 112. The encoded signal is modulated using a PN Code which here is labelled Pilot-PN_2 114 before eventually being transmitted through an antenna 118 to the mobile station 130. The first BTS 100 encodes a signal, which appears as a first interference signal to the mobile station 130, at ENCODER-1 102. This encoded signal is modulated using a PN Code Pilot-PN_1 104 before eventually being transmitted through an antenna 108. The third BTS 120 encodes a signal, which appears as a second interference signal to the mobile station 130, at ENCODER-3 122. This encoded signal is modulated using a PN Code which here is labelled Pilot-PN_3 124 before eventually being transmitted through an antenna 128. All three signals transmitted by antennas 108, 118, and 128 are received by the mobile station 130 at the receiver front-end 134 through antenna 132. The received signal is then passed to a decoder 138 for extraction of the channel to be recovered. The received signal is also passed on to a first correlator 140, a second correlator 142, and a third correlator 144. The correlators of figure 1, perform sub-operations corresponding to multiplication, summation, and absolute-value-squared, effectively performing an operation corresponding to taking an inner product of two inputs. The first correlator 140 performs a correlation between the received signal and the PN code Pilot-PN_1, which was used to modulate the signal appearing to the mobile as the first interference signal, and outputs an interference power I_1. The second correlator 142 performs a correlation between the signal and the PN code Pilot-PN_2, which was used to modulate the signal whose quality is to be measured, and outputs a signal power C. The third correlator 144 performs a correlation between the received signal and the PN code Pilot-PN_3, which was used to modulate the signal appearing to the mobile as the second interference signal, and outputs an interference power I_2. A calculating operation 150 computes the CIR which in this case is simply $C/(I_1+I_2)$.

In general, this approach can be applied to M base transceiver stations. Let BTS_i ($1 \le i \le M$) be the M adjacent base transceiver stations, E_i be the corresponding energy from the i^{th} base station that is measured at the mobile station 130, let S be the combined total signal energy received by the mobile at receiver front-end 134, and let BTS_2 be the base transceiver station whose associated CIR is to be measured, then

$$C = \max_{1 \le i \le M}(S \cdot PN_i) = E_2 \cdot N, \text{ and}$$

$$I = \sum_{i \ne 2}(S \cdot PN_i) = N \cdot \sum_{i \ne 2} E_i.$$

In these equations C and I are energies although for the purposes of determining the ratio C/I, either energy or power may be used. Since the pilot header is composed of two identical OFDM symbols, the CIR calculation process can be based on the average over the two symbols, thus reducing noise. These methods, however, fail to work if the channel is a multi-path fading channel and/or mobility speed is high. One solution is to insert more pilots to improve the measurement quality, however, this introduces overhead which significantly reduces spectral efficiency. For example, in 2G and 3G wireless systems, the pilot overhead is about 20-35%, and the pilot design for these systems is not suitable for fast channel quality measurement. This is the case because fundamentally the accuracy of the channel quality measurement is limited by the Cramer-Rao lower bound, which implies that the accuracy of channel measurement can be gained only at the expense of more pilot overhead (either in time or in power).

As an example of this trade-off, in a proposed MIMO-OFDM system, a pilot header is transmitted every OFDM frame in 10ms (15 slots). To facilitate adaptive modulation in the mobility case, a CIR estimation must be fed back to the BTS

every 2ms (3 slots). Therefore, CIR measurement based on a pilot header can not provide accurate instantaneous channel quality information. If the actual CIR does not change significantly during that 10 ms, then by measuring the energy of the pilots, one may roughly track the CIR. However, by doing so, the accuracy may diminish towards the end of the slot, as the assumption that the interference is a constant becomes more and more inaccurate.

The above discussed channel quality measurement is for adaptive coding and modulation, and does not in any way relate to channel estimation.

Channel quality measurement is a different concept from channel estimation. Channel quality measurement is performed to measure the channel quality so that proper coding and modulation set can be chosen. Channel estimation is performed to estimate the channel response so that coherent detection can be implemented.

In some wireless communication systems that employ Orthogonal Frequency Division Multiplexing (OFDM), a transmitter transmits data symbols to a receiver as OFDM frames in a MIMO (multiple input, multiple output) context. One of the key advantages of MIMO-OFDM systems is its ability to deliver high-speed data over a multi-path fading channel, by using higher QAM size, water pouring and/or adaptive modulation. In the MIMO-OFDM system, there are two major design challenges: (1) To combat high Doppler spread and fast fading due to high speed mobility (2) To provide a common fast signalling channel to realize fast physical and MAC layer adaptation signalling. To solve the mobility problem, a pilot channel is commonly used in OFDM design; such a pilot channel can be optimized by using the scattered (in time and frequency) pilot pattern. The common fast signalling channel design must

be sufficiently reliable to allow most of mobiles to detect the signalling, which introduces a significant amount of system and spectral overhead to sustain the signalling throughput. In the conventional OFDM design scattered pilot and fast signalling channel are arranged as separate overhead channels.

The phase and amplitude of the data symbols may be altered during propagation along a channel, due to the impairment of the channel. The channel response may vary with time and frequency. In order to allow the receiver to estimate the channel response, pilot symbols are scattered among the data symbols within the OFDM frame. The receiver compares the values of the received pilot symbols with the known transmitted values of the pilot symbols, estimates the channel response at the frequencies and times of the pilot symbols, and interpolates the estimated channel responses to estimate the channel response at the frequencies and times of the data symbols.

Transmit Parameter Signalling (TPS) symbols are also transmitted with the data symbols. The TPS symbols are transmitted over specified sub-carriers within the OFDM frame, and are used to provide common signalling channels to allow fast physical and media access control layer adaptation signalling.

Both the pilot symbols and the TPS symbols are overhead, in that they do not carry data. In order to improve the data rate of an OFDM communication system, the overhead within the OFDM frames should be minimized. The minimization of overhead is particularly important in Multiple-Input Multiple-Output (MIMO) OFDM systems. In a MIMO OFDM system having M transmitting antennae and N receiving antennae, the signal will propagate over M x N channels and there may be up to M sets of pilot symbols in the overhead. An example of an

OFDM frame format with dedicated TPS and pilot channels is shown in Figure 7 for the single input, single output case. The horizontal axis 704 shows a circle representing the frequency of each of a plurality of OFDM sub-carriers. The vertical axis 706 is time, with each row representing an OFDM symbol. A set of OFDM symbols constitutes an OFDM frame. In this example, the pilot channel is transmitted in a scattered manner, with the pilot symbols being transmitted every third sub-carrier, and for each sub-carrier every sixth frame. Thus, the first sub-carrier 700 has pilot symbols 701 in the first, seventh (and so on) OFDM symbols. The fourth sub-carrier 702 has pilot symbols 705 in the fourth, tenth (and so on) OFDM symbols. In addition, the third, ninth, 15^{th}, and 21^{st} sub-carriers of every OFDM symbol are used to transmit TPS symbols, collectively indicated at 708. The remaining capacity is used for traffic.

SUMMARY OF THE INVENTION

One embodiment of the invention provides a simple accurate and robust channel quality measurement method with broad applications such as UMTS and 3G wireless system evolution. Advantageously a channel quality indicator (CQI) is measured indirectly, simply, and accurately, and is independent of the mobile speed, independent of multi-path channel characteristics, and avoids Walsh Code Coherent Loss. The CQI is a measure of the overall quality of the channel, not just one factor, such as CIR. In addition the method is easy to implement, as it does not require any additional coding, such as PN codes used in CIR measurement.

According to one broad aspect, a channel quality measurement apparatus is provided which is adapted to measure a quality of a channel over which has been transmitted a sequence of symbols produced by encoding and constellation mapping a

source data element sequence. The apparatus has a symbol de-mapper, receiving as input a sequence of received symbols over the channel whose quality is to be measured, the symbol de-mapper being adapted to perform symbol de-mapping on said sequence of received symbols to produce a sequence of soft data element decisions. There is a soft decoder, receiving as input the sequence of soft data element decisions produced by the symbol de-mapper, the soft decoder being adapted to decode the sequence of soft data element decisions to produce a decoded output sequence. An encoder receives as input the decoded output sequence produced by the soft decoder, said encoder being adapted to re-encode the decoded output sequence with an identical code to a code used in encoding the source data element sequence to produce a re-encoded output sequence. Finally, a correlator, receives as input the sequence of soft data element decisions produced by the de-mapper, and the re-encoded output sequence produced by the encoder, said correlator being adapted to produce a channel quality indicator output by determining a correlation between the sequence of soft data element decisions and the re-encoded output sequence.

In some embodiments, the symbol de-mapper is adapted to perform QPSK symbol de-mapping.

In some embodiments, the symbol de-mapper is adapted to perform Euclidean distance conditional LLR symbol de-mapping.

Another broad aspect of the invention provides a method of measuring channel quality of a channel over which has been transmitted a sequence of symbols produced by encoding and constellation mapping a source data element sequence. The method involves receiving a sequence of received symbols over the channel whose quality is to be measured, symbol de-mapping said sequence of received symbols to produce a sequence of soft

data element decisions, decoding said sequence of soft data element decisions to produce a decoded output sequence, de-encoding said decoded output sequence to produce a re-encoded output sequence using a code identical to a code used in encoding the source data element sequence, and correlating said re-encoded output sequence, and said sequence of soft data element decisions to produce a channel quality indicator output.

In some embodiments, the method is applied to measure an OFDM channel quality.

Another broad aspect of the invention provides a communication system having a transmitter adapted to transmit a sequence of symbols produced by encoding and constellation mapping a source data element sequence over a channel; and a receiver having a) a symbol de-mapper, receiving as input a sequence of received symbols over the channel, said symbol de-mapper being adapted to perform symbol de-mapping on said sequence of received symbols to produce a sequence of soft data element decisions; b) a soft decoder, receiving as input the sequence of soft data element decisions produced by the symbol de-mapper, said soft decoder being adapted to decode the sequence of soft data element decisions to produce a decoded output sequence; c) an encoder, receiving as input the decoded output sequence produced by the soft decoder, said encoder being adapted to re-encode the decoded output sequence with an identical code to a code used in encoding the source data element sequence to produce a re-encoded output sequence; and d) a correlator, receiving as input the sequence of soft data element decisions produced by the de-mapper, and the re-encoded output sequence produced by the encoder, said correlator being adapted to produce a channel quality indicator output by determining a correlation between the sequence of soft data element decisions and the re-encoded output sequence.

The receiver is adapted to feed the channel quality indicator back to the transmitter, and the transmitter is adapted to use said channel quality indicator to determine and apply an appropriate coding rate and modulation to the source data element sequence.

Another broad aspect of the invention provides a method of adaptive modulation and coding which involves transmitting over a channel a sequence of symbols produced by encoding and constellation mapping a source data element sequence, receiving a sequence of received symbols over the channel, symbol de-mapping said sequence of received symbols to produce a sequence of soft data element decisions, decoding said sequence of soft data element decisions to produce a decoded output sequence, re-encoding said decoded output sequence to produce a re-encoded output sequence using a code identical to a code used in encoding the source data element sequence, correlating said re-encoded output sequence, and said sequence of soft data element decisions to produce a channel quality indicator output, transmitting the channel quality indicator, and using said channel quality indicator to determine and apply an appropriate coding rate and modulation to the source data element sequence.

Yet another broad aspect of the invention provides a method of determining a channel quality comprising correlating a soft data element decision sequence with a second data element sequence, the second data element sequence being produced by decoding the soft data element decision sequence to produce a decoded sequence and then re-encoding the decoded sequence.

Another broad aspect of the invention provides a method which involves applying forward error coding to a signalling message to generate a coded fast signalling message,

MPSK mapping the coded signalling message to produce an MPSK mapped coded signalling message, mapping the MPSK mapped coded signalling message onto a plurality of sub-carriers within an OFDM frame comprising a plurality of OFDM symbols, encoding symbols of the MPSK mapped coded signalling message using Differential Space-Time Block Coding (D-STBC) in a time direction to generate encoded symbols, and transmitting the encoded symbols on a plurality of transmit antennas, with the encoded symbols being transmitted at an increased power level relative to other symbols within the OFDM frame as a function of channel conditions.

In some embodiments, the encoded symbols are transmitted in a scattered pattern.

In some embodiments, transmitting the encoded symbols on a plurality of antennas involves: on a selected sub-carrier, each antenna transmitting a respective plurality N of encoded symbols over N consecutive OFDM symbols, where N is the number of antennas used to transmit, for a total of NxN transmitted encoded symbols, the NxN symbols being obtained from D-STBC encoding L symbols of the MPSK mapped coded signalling stream, where L,N determine an STBC code rate.

In some embodiments, the method further involves transmitting a set of pilot sub-carriers in at least one OFDM symbol, and using the pilot sub-carriers as a reference for a first set of D-STBC encoded symbols transmitted during subsequent OFDM symbols.

In some embodiments, transmitting a set of pilot sub-carriers in at least one OFDM frame involves transmitting a plurality of pilots on each antenna on a respective disjoint plurality of sub-carriers.

In some embodiments, each disjoint plurality of sub-carriers comprises a set of sub-carriers each separated by N-1 sub-carriers, where N is the number of antennas.

In some embodiments, pilot sub-carriers are transmitted for a number of consecutive OFDM frames equal to the number of transmit antennas.

An OFDM transmitter adapted to implement any of the above methods is also provided.

Another broad aspect of the invention provides a receiving method which involves receiving at at least one antenna an OFDM signal containing received D-STBC coded MPSK mapped coded signalling message symbols, recovering received signalling message symbols from the OFDM signal(s), re-encoding, MPSK mapping and D-STBC coding the received coded signalling message symbols to produce re-encoded D-STBC coded MPSK mapped coded signalling message symbols, and determining a channel estimate by comparing the received D-STBC coded mapped coded signalling message symbols with the re-encoded D-STBC coded MPSK mapped coded signalling message symbols.

In some embodiments, a channel estimate is determined for each location (in time,frequency) in the OFDM signal containing D-STBC coded MPSK mapped coded signalling message symbols. The method further involves interpolating to get a channel estimate for remaining each location (in time, frequency) in the OFDM signal.

In some embodiments, the method further involves receiving pilot symbols which are not D-STBC encoded which are used as a reference for a first D-STBC block of D-STBC coded MPSK mapped coded signalling message symbols.

An OFDM receiver adapted to implement any of the above methods is also provided.

An article of manufacture comprising a computer-readable storage medium is also provided, the computer-readable storage medium including instructions for implementing any of the above summarized methods.

Another broad aspect of the invention provides a method of generating pilot symbols from an Orthogonal Frequency Division Multiplexing (OFDM) frame received at an OFDM receiver, the OFDM frame containing an encoded fast signalling message in the form of encoded symbols within the OFDM frame. The method involves processing the encoded symbols based in a scattered pilot pattern to recover the encoded fast signalling message, re-encoding the fast signalling message so as to generate pilot symbols in the scattered pattern and recovering a channel response for the encoded symbols using decision feedback.

In some embodiments, the fast signalling message is examined to see if the current transmission contains content for the OFDM receiver. Only if this is true is the channel response computation process continued for the current transmission.

In some embodiments, processing the encoded symbols involves differentially decoding the encoded symbols using Differential Space-Time Block Coding (D-STBC) decoding to recover the encoded fast signalling message, applying Forward Error Correction decoding to the encoded fast signalling message to recover a fast signalling message, analyzing the fast signalling message to determine whether it includes a desired user identification and if the fast signalling message includes the desired user identification, re-encoding the fast signalling message using Forward Error Correction coding to

generate the encoded fast signalling message, and re-encoding the encoded fast signalling message using D-STBC.

Another broad aspect of the invention provides a transmitter adapted to combine pilot and transmission parameter signalling on a single overhead channel within an OFDM signal.

In some embodiments, a set of transmission parameter signalling symbols are transmitted on the overhead channel with strong encoding such that at a receiver, they can be decoded accurately, re-encoded, and the re-encoded symbols treated as known pilot symbols which can then be used for channel estimation.

Another broad aspect of the invention provides a receiver adapted to process the combined single overhead channel produced by the above summarized transmitter. The receiver is adapted to decode a received signal containing the encoded transmission parameter signalling symbols as modified by a channel, re-encode the decoded symbols to produce known pilot symbols, compare received symbols with the known pilot symbols to produce a channel estimate.

Other aspects and features of the present invention will become apparent to those of ordinary skill in the art upon review of the following description of specific embodiments of the invention in conjunction with the accompanying figures.

BRIEF DESCRIPTION OF THE DRAWINGS

The invention will now be described in greater detail with reference to the accompanying diagrams, in which:

Figure 1 is a diagram of a standard carrier to interference ratio (CIR) estimator using a known channel quality measurement technique;

Figure 2 is a diagram of a channel quality indicator (CQI) estimator constructed according to an embodiment of the invention;

Figure 3 is a graph showing a QAM constellation to illustrate QPSK de-mapping according to an embodiment of the invention;

Figure 4 is a graph showing simulation results of CQI versus SNR for different Doppler frequencies;

Figure 5 is graph showing statistical results of CQI measurements;

Figure 6 is a graph showing a CDF of SNR measurement error based on the CQI;

Figure 7 is a diagram of OFDM symbol allocation for dedicated pilot and TPS channels;

Figure 8 is a block diagram of an OFDM system employing combined TPS and pilot signalling in a single overhead channel provided by an embodiment of the invention;

Figure 9 is an OFDM symbol allocation diagram showing time and frequency differentials;

Figure 10 is an example of an OFDM symbol allocation diagram showing pilot and TPS symbol locations; and

Figures 11 and 12 are example performance results for the system of Figure 8.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

According to one embodiment of the invention, a measurement of the quality of the received signal is obtained by measuring a value representative of the average distance

between the received signal and the reference signal constellation. In general the poorer the channel, the more scattered and random is the received signal on the reference signal constellation, and therefore the larger the average distance between the signal and its closest constellation reference point.

In some implementations, the purpose of channel quality measurement, as was the case for C/I estimation, is for a successful coding rate and modulation assignment. A "successful" assignment here is one which achieves desired performance characteristics. In accordance with this purpose, a new channel quality measurement referred to herein as the "Channel Quality Indicator" (CQI) is provided. The CQI provides an overall assessment of the quality of the channel, including the effects of interference, multi-path fading, and Doppler spread.

In developing the CQI, a soft output from a de-mapping function is used to obtain a measurement of channel quality, since the amplitude of the soft output can be used as an indication of the confidence of the signal. If the channel quality is high, the soft output value will be high, and vice versa. All the channel impairments will be reflected in such an indicator, independent of their source and character. This has been demonstrated by simulation results, which show that such an indicator is invariant to the interference, multi-path fading and Doppler spread.

The preferred embodiment presented is based on an MIMO-OFDM frame structure in which a QAM constellation is employed, and provides an indirect channel quality measurement approach based on soft QAM demodulation and de-mapping. However, more generally, embodiments of the invention provide for any frame structure which employs a method of modulation

and mapping having an associated reference symbol constellation which can be used in soft demodulation and de-mapping such as PSK (phase shift keying) and PAM (pulse amplitude modulation) to name a few examples.

Referring to figure 2, a preferred embodiment of the invention will now be described. It is assumed for the purpose of this example that a signal from a second base transceiver station 210 is a desired signal whose associated channel quality is to be measured by a mobile station 230, and that signals from two other (first and third) base stations 200, and 220, can be considered to be noise by mobile station 230. There may be other sources of noise as well, and the channel may introduce distortions such as multi-path fades, residual Doppler shifts, and thermal white noise. The second BTS 210 encodes an input sequence 213 (assumed to be a sequence of bits, but more generally a sequence of data elements) at ENCODER-2 212 to produce an encoded bit sequence. The encoded bit sequence contains redundancy which allows some error detection/correction at the receiver. The encoded bit sequence is then mapped to constellation points with symbol mapper 214. These constellation points are modulated and transmitted as a signal whose associated channel quality is to be measured. The signal is transmitted through an antenna 218 to a mobile station 230. The modulation type (and associated constellation) and type of coding employed by ENCODER-2 212 are both adaptively selected as a function of a channel quality indicator fed back from the mobile station 230.

The first BTS 200 encodes with ENCODER-1 202 and maps with symbol mapper 204 to produce a signal, which appears as a first interference signal to the mobile station 230. This signal is transmitted through an antenna 208. The third BTS 220 encodes with ENCODER-3 222 and maps with symbol mapper 224 to produce a signal which appears as a second interference

signal to the mobile station 230. This signal is transmitted through an antenna 228. All three channels transmitted by antennas 208, 218, and 228 are received by the mobile station 230 at the receiver front-end 234 through antenna 232, although in this example, the signal from the second base transceiver station 210 is the desired signal. According to the preferred embodiment, the received signal is then passed to a symbol de-mapper 236. The symbol de-mapper 236 takes raw symbol data from the receiver front end 234 and de-maps the raw symbol data taking into account the known signal constellation used at the transmitting base station 210 to produce a soft bit decision sequence. The de-mapped symbols (soft bit decisions) inherently constitute a representation of confidence, and are used as inputs to a soft decoder 238. The symbol de-mapper 236 outputs a de-mapped output signal at output 237 both to the soft decoder 238 and to a correlator 250. The soft decoder 238 performs soft decoding on the de-mapped output signal and outputs a soft decoded output signal to an encoder 240. The soft decoded output is also output at 239 as a receiver output, this being the best available estimate at the receiver of input sequence 213. Alternatively, a different receiver structure may be used to generate a receiver output. The encoder 240 re-encodes the output of the soft decoder to produce an encoded output signal and outputs this encoded output signal from output 242 to a correlator 250. The same encoding is used as was employed at ENCODER-2 212 of the base station 210. Assuming proper decoding and re-encoding, the output of the encoder 240 is the same as the encoded sequence produced by the encoder 212 at the base transceiver station 210. The correlator 250 correlates the re-encoded sequence from the encoder output 242, with the de-mapped output signal (soft bit decision sequence) from the symbol de-mapper output 237. The correlator 250 outputs this correlation as a channel quality indicator (CQI). The higher this correlation, the closer the de-mapped symbols

are on average to the transmitted constellation symbols and as such the higher the channel quality. In the illustrated example, correlator 250 multiplies the re-encoded bit sequence 242 with the soft bit decision sequence with multiplier 251. These are summed with summer 252, and then the square absolute value is taken as indicated at 253. Other methods of correlating may be employed.

In one example implementation, the symbol de-mapper 236, takes the input from the receiver front end 234, and performs de-mapping based on Euclidean distance. The preferred embodiment is described in the context of QPSK de-mapping, which is a special case of PSK de-mapping. Generally for PSK modulation, there are two types of de-mapping methods based on whether or not the PSK signals have been normalized. For coherent de-mapping, since the exact reference constellations are known, the optimum de-mapping is based on Euclidean distance; while for non-coherent de-mapping, which is often the case when differential encoding is used, de-mapping can only be based on angle. The de-mapping-based-on-angle method is a sub-optimum one, as it ignores the information carried in the amplitude of a signal. As a special case of PSK de-mapping, QPSK de-mapping does not depend upon signal normalization. As is the case in de-mapping higher QAM signals, QPSK de-mapping is based on an LLR (logarithm of likelihood ratio) and in this example, as described with reference to figure 3, uses Euclidean distance. The constellation depicted in figure 3 is a QPSK constellation with Grey mapping. Corresponding to the bit sequences 00, 01, 10 and 11 are constellation points S_0, S_1, S_2, and S_3 respectively, whose co-ordinates are (x_0, y_0), (x_1, y_1), (x_2, y_2), and (x_3, y_3) respectively. Point (x, y) represents the signal input from the receiver front end 234. The soft de-

mapped bits $b_1 b_2$ using Euclidean distance LLR can be expressed as:

$$b_1 = \log \frac{e^{-\left((x-x_2)^2+(y-y_2)^2\right)/2\sigma^2} + e^{-\left((x-x_3)^2+(y-y_3)^2\right)/2\sigma^2}}{e^{-\left((x-x_0)^2+(y-y_0)^2\right)/2\sigma^2} + e^{-\left((x-x_1)^2+(y-y_1)^2\right)/2\sigma^2}}$$

$$b_2 = \log \frac{e^{-\left((x-x_1)^2+(y-y_1)^2\right)/2\sigma^2} + e^{-\left((x-x_3)^2+(y-y_3)^2\right)/2\sigma^2}}{e^{-\left((x-x_0)^2+(y-y_0)^2\right)/2\sigma^2} + e^{-\left((x-x_2)^2+(y-y_2)^2\right)/2\sigma^2}},$$

where $\sigma^2 = 2EN_o$, and E is the energy of per QPSK symbol.

The calculation of bit b_1 can be simplified. Since the four QPSK constellation points have equal distance to the origin $(0,0)$:

$$x_0^2 + y_0^2 = x_1^2 + y_1^2 = x_2^2 + y_2^2 = x_3^2 + y_3^2.$$

Then b_1 simplifies to:

$$b_1 = \log \frac{e^{-\left((x-x_2)^2+(y-y_2)^2\right)/2\sigma^2} + e^{-\left((x-x_3)^2+(y-y_3)^2\right)/2\sigma^2}}{e^{-\left((x-x_0)^2+(y-y_0)^2\right)/2\sigma^2} + e^{-\left((x-x_1)^2+(y-y_1)^2\right)/2\sigma^2}}$$

$$= \log \frac{e^{(xx_2+yy_2)/\sigma^2} + e^{(xx_3+yy_3)/\sigma^2}}{e^{(xx_0+yy_0)/\sigma^2} + e^{(xx_1+yy_1)/\sigma^2}}$$

$$= \log \frac{e^{(xx_3+yy_3)/\sigma^2}\left(1+e^{(xx_2+yy_2)-(xx_3+yy_3)/\sigma^2}\right)}{e^{(xx_1+yy_1)/\sigma^2}\left(1+e^{(xx_0+yy_0)-(xx_1+yy_1)/\sigma^2}\right)}$$

Since $x_0 = x_1$ and $x_2 = x_3$:

$$b_1 = \log \frac{e^{(xx_3+yy_3)/\sigma^2}\left(1+e^{y(y_2-y_3)/\sigma^2}\right)}{e^{(xx_1+yy_1)/\sigma^2}\left(1+e^{y(y_0-y_1)/\sigma^2}\right)}$$

Let D be the vertical distance in the I-Q plot between S_0 and S_1, and between S_2 and S_3. Therefore $y_0 - y_1 = y_2 - y_3 = D$, and:

$$b_1 = \log \frac{e^{(xx_3+yy_3)/\sigma^2}}{e^{(xx_1+yy_1)/\sigma^2}}$$

$$= \frac{1}{\sigma^2}\log\left(e^{x(x_3-x_1)+y(y_3-y_1)}\right)$$

Because of the symmetry of the constellation $x_3 - x_1 = -D$. Since $y_1 = y_3$, b_1 can be expressed as:

$$b_1 = -\frac{D}{\sigma^2}x$$

Similarly, b_2 is expressed as:

$$b_2 = -\frac{D}{\sigma^2}y$$

If the noise is fixed, then the QPSK de-mapping algorithm can be simplified further to:

$$b_1 = -x$$

$$b_2 = -y\,,$$

This is equivalent to two BPSK signals and is very easy to compute.

In STBC (Space-Time Block Coding), the combined QPSK signal x is normalized by the factor $\delta^2 = |h_{11}|^2 + |h_{21}|^2 + |h_{12}|^2 + |h_{22}|^2$

where $h_{n,m}$ are elements of an MIMO (Multiple Input Multiple Output) channel matrix. Suppose the noise variances of the four channels are the same, i.e., σ^2, then the noise power becomes $(\sigma/\delta)^2$. Thus b_1 with STBC is

$$b_1 = -\frac{D}{\left(\sigma/\delta\right)^2}\left(\frac{x}{\delta^2}\right)$$

$$= -\frac{D}{\sigma^2}x$$

This verifies therefore that QPSK in STBC de-mapping is not affected by different scaling factors used in normalization. The conditional LLR soft de-mapped bits $b_1 b_2$ are output to the soft decoder 238 which uses the de-mapped bits, and takes into account the data stream history information, the encoding algorithm which was used in Encoder-2 212, to make a best estimate of the original unencoded code word. This best estimate which is output from the soft decoder 238 is re-encoded by encoder 240 using the same encoding algorithm as encoder-2 212. The re-encoded code word is output from encoder output 242, to the correlator 250. The correlator 250 correlates the conditional LLR output from output 237 of the symbol de-mapper 236, with the re-encoded code word output from output 242 of the encoder 240. The act of correlation projects the conditional LLR onto the re-encoded code word, the result of which is an inner product output which is used as the Channel Quality Indicator (CQI).

Advantageously the CQI, because it is a measure of the correlation between the symbol de-mapper output and the re-encoded sequence, indicates the channel distortion. The use of the likelihood value relies neither on the code type (block code, convolutional code, or turbo code), nor on the decoding

method (hard or soft), and does not distinguish where the interference originates, e.g., neighboring-cell interference, white thermal noise, or residual Doppler shift. The CQI uses all the information available for the estimation, not only the values of the de-mapped output, but the likelihood of being a code word as well, which is much more accurate than measuring soft output value alone, especially when the code rate is low. In figure 4, simulation results are shown in a graph of normalized CQI versus SNR for different Doppler frequencies for the Bi-orthogonal code (16,5). In figure 5 statistical SNR measurement error results are shown, and in figure 6, simulation results are shown in a CDF of SNR measurement error based on the CQI. These graphs show that for a given BER, the CQI is relatively invariant with respect to various Doppler frequencies and different channel models. This means that conversely, irrespective of channel conditions, the CQI can be used to provide a consistent representation of BER, and as such using the CQI to perform adaptive coding and modulation decisions, a desired BER can be achieved. This is accomplished by feeding back the CQI to the transmitter whose signal is associated with the channel whose quality is to be measured. Based on the CQI and the desired performance, the transmitter determines and applies the appropriate coding rate and modulation.

Combined Pilot and TPS Channel

In the above embodiment, coded transmit data is used at a receiver to generate a channel quality indicator for use in making adaptive coding and modulation decisions. In another embodiment of the invention, a method is provided of combining pilot symbols with Transmit Parameter Signalling (TPS) symbols within an Orthogonal Frequency Division Multiplexing (OFDM) frame in such a manner that channel estimation can still be performed. The method may be implemented at a SISO (single-

input single-output) transmitter or implemented at a Multiple-Input Multiple-Output (MIMO) OFDM transmitter, and can be described broadly as four steps. First, a fast signalling message is forward error coding (FEC) encoded to generate a coded fast signalling message. Second, the coded fast signalling message is mapped onto symbols within the OFDM frame. Third, the symbols are encoded using Differential Space-Time Block Coding (D-STBC) to generate encoded symbols. The D-STBC coding is preferably applied in the time direction of the OFDM frame, as the channel response of the channel over which scattered pilot sub-carriers are transmitted will usually vary more rapidly with frequency direction than with time direction, and so differential decoding at the OFDM receiver is more likely to yield a better estimate of the channel response if the differential decoding is with respect to symbols distributed along the time direction. Fourth, the encoded symbols are transmitted in a scattered pilot pattern at an increased power level relative to other traffic data symbols within the OFDM frame. In some embodiments, the power level is only increased relative to other traffic data symbols if channel conditions are poor.

The method allows fast signalling messages to be used as pilot symbols, thereby reducing overhead within the OFDM frame.

A method of extracting pilot symbols from an OFDM frame in which the pilot symbols have been combined with TPS symbols, as described above, is also provided. The method is implemented at a MIMO OFDM receiver when an OFDM frame containing encoded symbols is received at the OFDM receiver, and can be described broadly as eight steps. First, the OFDM receiver recovers the encoded symbols based on the scattered pattern to recover the D-STBC blocks. Second, the OFDM receiver differentially decodes the recovered D-STBC blocks

using D-STBC decoding to recover the FEC encoded fast signalling message. Third, the OFDM receiver applies FEC decoding to the FEC encoded fast signalling message to recover the fast signalling message. Fourth, the OFDM receiver analyzes the fast signalling message to determine whether it includes a desired user identification. If the fast signalling message includes the desired user identification, then the OFDM receiver knows that the current TPS frame contains data for the user and continues processing the OFDM frame. As a fifth step, the OFDM receiver re-encodes the fast signalling message using FEC coding. Sixth, the OFDM receiver re-encodes the encoded fast signalling message using D-STBC encoding. If the fast signalling message does not include the receiver's user identification, then power can be saved by not proceeding to conduct the rest of the channel estimation steps.

Now the TPS symbols having been D-STBC re-encoded can be used as pilots. A channel response for the D-STBC encoded symbol can be obtained by comparing the known transmitted pilots (re-encoded TPS data) with the received signals. A channel response is obtained for each TPS insertion point. The channel responses thus determined can then be used to interpolate a channel response for every traffic data symbol, at all times and frequencies, within the OFDM frame. Preferably, this is done by performing a 2-dimensional interpolation (in time direction and frequency direction) to generate channel estimate for some points where TPS were not inserted. This is followed by an interpolation in frequency to generate a channel estimate for every sub-carrier of OFDM symbols containing TPS data. In some embodiments, every OFDM symbol contains some TPS insertion points and as such this completes the interpolation process. In other embodiments, there are some OFDM symbols which do not have any TPS insertion points. To get channel estimates for these OFDM symbols, an

interpolation in time of the previously computed channel estimates is performed. In high mobility applications, TPS should be included in every OFDM symbol avoiding the need for this last interpolation in time step.

A fast algorithm may be applied at the OFDM receiver when computing a Discrete Fourier Transform based on the scattered pattern in order to extract the combined pilot and fast signalling message. This reduces power consumption at the OFDM receiver.

The invention has been described with respect to a MIMO-OFDM communication system. The invention may also be used in a single transmitter OFDM communication system, but will be of less advantage as the number of pilot symbols transmitted as overhead is more manageable than in MIMO OFDM communication systems.

The method of combining pilot symbols with the TPS channels and the method of extracting pilot symbols are preferably implemented on an OFDM transmitter and on an OFDM receiver respectively in the form of software instructions readable by a digital signal processor. Alternatively, the methods may be implemented as logic circuitry within an integrated circuit. More generally, the methods may be implemented by any computing apparatus containing logic for executing the described functionality. The computing apparatus which implements the methods may be a single processor, more than one processor, or a component of a larger processor. The logic may comprise external instructions stored on a computer-readable medium, or may comprise internal circuitry.

One of the constraints of conventional STBC is the need for accurate knowledge of channel information. In order to eliminate the requirements for channel knowledge and pilot

symbol transmission, D-STBC is preferable for high mobility application.

A detailed example will now be provided for the case where a 2-input, 2-output system is being employed, although the technology is applicable to arbitrary numbers of antennas. also, for this example, an OFDM symbol having 25 sub-carriers is assumed, although any number of sub-carriers may be employed. This example is assumed to operate on with frames of 16 OFDM symbols, but more generally any length of frame may be employed.

A preferred D-STBC scheme is shown in Figure 8 and described in detail below. To design the D-STBC for MIMO-OFDM, there are 3 major issues to be addressed.

1. Differential direction,
2. Data protection,
3. Initialization/reset.

Differential Direction

One of the critical assumptions for any differential encoding is that the channel variation between two coded symbols should be *sufficiently small*. For the time-frequency structure of the OFDM signal as shown in Figure 9, the channel variation along the frequency axis represents the multi-path channel induced frequency selectivity, the channel variation along the time axis represents the temporal fading variation. The differential encoding direction should be optimized.

Differential in frequency is limited by the channel coherence bandwidth determined by the multi-path delay spread. The phase shift between two adjacent pilots could be very large, for example, for the ITU Vehicular A channel, if the two pilot blocks are 16 bins apart, then the phase shift of the channel between the two positions can be as high as π, which

makes differential decoding impossible. To solve this problem, the span of pilots in the frequency domain must be reduced. However, this will further increase the pilot overhead.

Differential in time is limited by Doppler frequency caused by high-speed mobility. For practical channel models, we can assume that the channel remains approximately the same along several OFDM symbols. The channel variation along the time direction varies much slower than along the frequency direction, therefore, D-STBC should preferably be encoded along time direction. According to a preferred embodiment of the invention, due to the STBC structure, a pair of the STBC encoded TPS symbols are allocated on the same frequency index (sub-carrier) of two adjacent OFDM symbols. The two possible differentials are shown in Figure 9. Differential in time encoding is generally indicated by 900 and differential in frequency encoding is generally indicated by 902.

Data Protection

FEC encoding is preferably applied to TPS data, since the decoding of the TPS data is critical for configuring the receiver to detect the traffic data correctly and for the correct re-encoding of the TPS data so as to allow an accurate decision feedback to reliably convert the TPS into a scattered pilot. A (32, 6) Hadamard code might for example be used. However, the code selection is not limited to this code alone.

Initialization and Reset

D-STBC relies on two consecutively received code blocks to decode the current block of data. Since the OFDM header may not employ D-STBC for the reason of frequency offset and sampling frequency estimation etc., the first received D-STBC block does not have any previous blocks to do the differential processing. This means that the first block of TPS cannot carry any signaling information. To solve this

problem, preferably pilot channel OFDM symbols are periodically inserted in the OFDM symbols. An example of this is shown in Figure 10 where pilot symbols are inserted in every sub-carrier periodically, for example 2 pilot channel OFDM symbols for every 20 OFDM symbols. The pilot symbols transmitted on the pilot channel OFDM symbols are preferably sent only by one antenna at a time for a given frequency. For example, in a two antenna system, the pilot symbols may alternate in frequency between the first and second antenna. This is shown in Figure 10 where two OFDM symbols 910,912 are used to transmit pilot symbols, and every odd sub-carrier is used for the first antenna, and every even sub-carrier is used for the second antenna. These pilot symbols may then be used as a reference for subsequent D-STBC symbols. For each antenna, interpolation can be performed to obtain pilot information for the intervening non-transmitted sub-carriers. Thus, interpolation is performed for the even sub-carriers for the first transmitter, and interpolation is performed for the odd sub-carriers for the second transmitter.

The channel information obtained from the pilot header is then used to decode the first blocks of TPS. Since the pilot header is transmitted periodically, the D-STBC encoder is also reset at the same frequency. After the first blocks of TPS are processed, the user has also obtained the first blocks of D-STBC references. In addition, the resetting of D-STBC encoder by periodic pilot headers prevents error propagation in the decision-feedback channel estimation process.

Figure 10 also shows the example locations of TPS symbols and of data symbols. In this example, the first two OFDM symbols 910,912 of every 20 symbol cycle contain pilot symbols as discussed above. The third through 20th frames contain TPS or data. A diamond lattice pattern is used for TPS

symbols, with every third sub-carrier containing TPS symbols, alternating between three sets of two TPS symbols on the first, seventh, thirteenth, nineteenth and twenty-fifth sub-carriers 914,916,918,920,922, and two sets of two TPS symbols on the fourth, tenth, sixteenth and twenty-second sub-carriers 924,925,926,928.

Unlike the pilot symbols transmitted in frames 910,912 which are transmitted by one antenna per sub-carrier, for each TPS symbol location shown in Figure 10 TPS data is transmitted all of the antennas, (i.e. by both antennas in our example). The TPS data transmitted on the two antennas collectively forms a common TPS channel.

Figure 11 shows TPS bit error rate versus SNR curves for various Doppler frequencies. As we can see from the figure, it is very robust to Doppler spread. Figure 12 shows the simulation results for traffic channel based on TPS assisted channel estimation. From this figure, it can be seen that the degradation due to TPS decoding error is negligible.

The details of the preferred D-STBC approach will now be explained. D-STBC involves the recursive computing of a transmission matrix. By "differential," it is meant the current transmitted D-STBC block is the matrix product operation between the previously transmitted D-STBC block and the current STBC block input.

As indicated previously, preferably TPS data is transmitted on two consecutive OFDM symbols for the same sub-carrier for a set of sub-carriers which may change from one set of two OFDM symbols to another set of two OFDM symbols. More generally, for a MIMO system with N antennas, TPS data is transmitted over N consecutive OFDM frames for the same sub-carrier. The transmission matrix is an NxN matrix that determines what to transmit on the N (consecutive OFDM frames) x N (number of antennas) available TPS symbol locations. For

the example being described in detail, N=2. The actual amount L of TPS data transmitted depends on the D-STBC code rate. For example, if there are four antennas, then a 4x4 STBC matrix is obtained from encoding three symbols from the MPSK mapped TPS signalling stream.

Referring to Figure 10, the first sub-carrier transmitted by both antennas will contain TPS data on the third, fourth, ninth, tenth, and 15^{th}, 16^{th} frames. The data will be both time and space differentially encoded meaning that there is information both in the difference between symbols sent at different times (differential time), and in the difference between symbols sent on different antennas (differential space).

The first and second pilot symbols 930 (frame 910) and 932 (frame 912) transmitted by the first antenna on the first sub-carrier and an interpolated value for the first pilot and second pilot symbols transmitted by the second antenna on the first sub-carrier collectively provide a reference for the first two TPS symbols 934,936 transmitted by the two antennas. Subsequent TPS symbols rely on previously transmitted TPS symbols as references.

Referring now to Figure 8, the forward error corrected TPS data to be transmitted on a given sub-carrier is indicated as a sequence $\{C_1, C_2 \ldots\}$ 950, assumed to be M-ary in nature. This is M-PSK mapped at 952. M-PSK symbols are then processed pairwise (for the 2x2 case) with a pair of M-PSK symbols at time i being referred to as $\{x_{1,i}, x_{2,i}\}$. Space time block coding produces a 2x2 STBC matrix $H_{x,i}$ 954 which contains $x_{1,i}$,$x_{2,i}$ in a first column and $-x_{2,i}^*$, $x_{1,i}^*$ in the second column. For the purpose of the TPS frames, the STBC block index i increments once every 2 OFDM symbols. A counter m will represent OFDM symbols with the m^{th} and $m+1^{th}$ OFDM symbol from transmitter STBC block index i, m=2i. In the Figure, the

output of the encoder at time i is identified as $H_{z,i}$, 956 with the output at time i-1 identified as $H_{z,i-1}$ stored in delay element 958. $H_{z,i}$ has the same structure as $H_{x,i}$. The following encoder equation can be obtained for the output as a function of the input:

$$H_{z,i} = \frac{1}{\sqrt{E_x}} H_{x,i} H_{z,i-1}$$

where $H_{z,i}$ is the D-STBC matrix at STBC block index i, $H_{x,i}$ is the STBC input matrix at STBC block index i, and E_x is the energy of each signal in $H_{x,i}$. The output $H_{z,i}$ is a 2x2 matrix having four elements with the first row of the elements being transmitted on one antenna 960, and second row of the elements being transmitted on the other antenna 962. For the example of Figure 10, the matrix $H_{z,i}$ is transmitted collectively by the two antennas during TPS symbol locations 934,936 of the first sub-carrier using the pilot symbols as the reference.

Referring again to Figure 8, at a single antenna receiver, the antenna receives a signal $Y_1 = y_1(m), y_1(m+1)$ at STBC block index i over two OFDM frames $m, m+1$ for each sub-carrier. This will be received on a single sub-carrier over two OFDM frames.

To understand D-STBC is to observe the following key equation which holds true for antenna 1:

$$\begin{aligned} \begin{bmatrix} y_1(m) \\ y_1(m+1) \end{bmatrix} &= H_{z,i} A_{1,i} \\ &= \frac{1}{\sqrt{E_x}} H_{x,i} H_{z,i-1} A_{1,i} \\ &\approx \frac{1}{\sqrt{E_x}} H_{x,i} \begin{bmatrix} y_1(m-2) \\ y_1(m-1) \end{bmatrix} \end{aligned}$$

where $y_1(m)$, $y_1(m+1)$ is the received signal over two OFDM frames for STBC block index i, $H_{x,i}$ is the STBC block input at STBC

block index i, E_x is the energy of signal elements in $H_{x,i}$, $A_{1,i}$ is the channel matrix for receive antenna 1 representing the channel response h_{11} from first transmit antenna to the receive antenna and h_{21} for the second transmit antenna to the receive antenna at STBC block index i, and $H_{z,i}$ is the transmitted D-STBC block signal at STBC block index i. D-STBC can only be applied to PSK modulation, and therefore, E_x is a fixed value. Also, $H_{z,i}$ takes the same format as $H_{x,i}$, i.e.,

$$H_{z,i} = \begin{bmatrix} z_{1,i} & z_{2,i} \\ -z_{2,i}^{*} & z_{1,i}^{*} \end{bmatrix}.$$

From the equation

$$\begin{bmatrix} y_1(m) \\ y_1(m+1) \end{bmatrix} \approx \frac{1}{\sqrt{E_x}} H_{x,i} \begin{bmatrix} y_1(m-2) \\ y_1(m-1) \end{bmatrix}$$

we can obtain $H_{x,i}$ from the four consecutively received signals $y_1(m-2)$, $y_1(m-1)$, $y_1(m)$, $y_1(m+1)$. Note that in the case of multiple receiver antennas, the same expression holds true for each antenna. Since D-STBC works on STBC blocks, it also has the same soft failure property as STBC, i.e., the system will not break down due to transmitting antennas failure - as long is still at least one antenna working. In addition, the code design for MIMO channel is in fact a task for STBC, and is irrelevant to D-STBC. Therefore, D-STBC can be easily expanded to the case with transmitter diversity of order more than 2.

Other System Design Considerations

Encoding

Although theoretically the differential encoding is after STBC encoding, (i.e. STBC matrix Hx,i is computed and then Hz,i is computed), in practice, these steps can be

reversed in order. The main advantage of reversing the order is that the STBC encoding process can be unified, which makes it very simple and easy to implement. To elaborate, we can calculate $z_{1,i}$ and $z_{2,i}$ from $x_{1,i}$ and $x_{2,i}$ first, then puncture or insert $z_{1,i}$ and $z_{2,i}$ into the data stream that are to be STBC encoded. The elements $z_{1,i}$ and $z_{2,i}$ can be calculated as follows:

$$z_{1,i} = \frac{1}{\sqrt{E}}\left(x_{1,i}z_{1,i-1} - x_{2,i}z_{2,i-1}{}^{*}\right)$$
$$z_{2,i} = \frac{1}{\sqrt{E}}\left(x_{1,i}z_{2,i-1} + x_{2,i}z_{1,i-1}{}^{*}\right)$$

The above equation is the only operation needed for D-STBC encoder, where no matrix operation is involved. One row of the resultant matrix $H_{z,i,}$, namely $z_{1,i}$, $z_{2,i}$ is transmitted by one antenna, and the other row, namely $-z_{2,i}^{*}$, $z_{1,i}^{*}$ is transmitted by the other antenna.

Decoding

The decoding of differentially encoded STBC code can be simplified into one step even simpler than STBC decoding itself, considering that there is no channel estimation is needed. Note that all the calculation here is carried out in the frequency domain, therefore, the relation between the transmitted signal and the channel is multiplication, rather than convolution.

Define:

- m : OFDM symbol index in time
- i : OFDM channel estimation index = 2m
- k : OFDM sub-carrier index
- $x_{1,i}$: first PSK symbol to form STBC block $H_{x,i}$
- $x_{2,i}$: second PSK symbol to form STBC block $H_{x,i}$
- $y_j(m)$: received signal at antenna $j=1,2$

The transmitted STBC coded signal (i.e., before the differential encoder) at time m and $m+1$ is:

$$\begin{bmatrix} x_{1,i} & x_{2,i} \\ -x_{2,i}^{*} & x_{1,i}^{*} \end{bmatrix},$$

where the column number is in space domain, while the row number is in time domain. Note the relationship hold true on a per sub-carrier basis.

With differential coding, the received signal at two receiving antennas for STBC block index can be expressed as follows for each sub-carrier, (sub-carrier index not shown) where again $m = 2i$:

$$\begin{bmatrix} y_1(m) \\ y_1(m+1) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} x_{1,i} & x_{2,i} \\ -x_{2,i}^{*} & x_{1}^{*} \end{bmatrix} \begin{bmatrix} y_1(m-2) \\ y_1(m-1) \end{bmatrix}$$

$$\begin{bmatrix} y_2(m) \\ y_2(m+1) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} x_{1,i} & x_{2,i} \\ -x_{2,i}^{*} & x_{1,i}^{*} \end{bmatrix} \begin{bmatrix} y_2(m-2) \\ y_2(m-1) \end{bmatrix}$$

From the above two equations, the maximum likelihood signals of $x_{1,i}$ and $x_{2,i}$ can be obtained as:

$$\begin{aligned} \tilde{x}_{1,i} &= y_1(m-2)^{*} y_1(m) + y_1(m-1) y_1(m+1)^{*} \\ &\quad + y_2(m-2)^{*} y_2(m) + y_2(m-1) y_2(m+1)^{*} \end{aligned}$$

$$\begin{aligned} \tilde{x}_{2,i} &= y_1(m-1)^{*} y_1(m) - y_1(m-2) y_1(m+1)^{*} \\ &\quad + y_2(m-1)^{*} y_2(m) - y_2(m-2) y_2(m+1)^{*} \end{aligned}$$

or in a matrix form:

$$\begin{aligned} \begin{bmatrix} \tilde{x}_{1,i} \\ \tilde{x}_{2,i} \end{bmatrix} &= \begin{bmatrix} y_1(m-2)^{*} & y_1(m-1) \\ y_1(m-1)^{*} & -y_1(m-2) \end{bmatrix} \begin{bmatrix} y_1(m) \\ y_1(m+1)^{*} \end{bmatrix} \\ &\quad + \begin{bmatrix} y_2(m-2)^{*} & y_2(m-1) \\ y_2(m-1)^{*} & -y_2(m-2) \end{bmatrix} \begin{bmatrix} y_2(m) \\ y_2(m+1)^{*} \end{bmatrix} \end{aligned}$$

It is the above matrix equation is depicted in block diagram form in the receiver path of Figure 8.

Channel estimation

Since the finally transmitted data are D-STBC encoded, channel parameters for each path can only be estimated through re-encoding the decoded data, after TPS have been successfully decoded. This decision-feedback approach is the key in how to make use of TPS as scattered pilots.

Suppose that after D-STBC re-encoding, we obtain $z_{1,i}$ and $z_{2,i}$, which correspond to $x_{1,i}$ and $x_{2,i}$, respectively, then from receiver antenna 1 we have

$$\begin{bmatrix} y_1(m) \\ y_1(m+1) \end{bmatrix} = \begin{bmatrix} z_{1,i} & z_{2,i} \\ -z_{2,i}^* & z_{1,i}^* \end{bmatrix} \begin{bmatrix} h_{11}(m) \\ h_{21}(m) \end{bmatrix}.$$

By solving the above equation, we get

$$\begin{bmatrix} h_{11}(m) \\ h_{21}(m) \end{bmatrix} = \frac{1}{\delta^2} \begin{bmatrix} z_{1,i}^* & -z_{2,i} \\ z_{2,i}^* & z_{1,i} \end{bmatrix} \begin{bmatrix} y_1(m) \\ y_1(m+1) \end{bmatrix},$$

where

$$\delta^2 = |z_{1,i}|^2 + |z_{2,i}|^2 .$$

In a similar way, we can estimate $h_{12}(m,k)$ and $h_{22}(m,k)$ from the signals received at receiver antenna 2:

$$\begin{bmatrix} h_{12}(m) \\ h_{22}(m) \end{bmatrix} = \frac{1}{\delta^2} \begin{bmatrix} z_{1,i}^* & -z_{1,i} \\ z_{2,i}^* & z_{2,i} \end{bmatrix} \begin{bmatrix} y_2(m) \\ y_2(m+1) \end{bmatrix}.$$

It needs to be noticed that for each STBC block, we can only obtain one set of channel information for the current time, with the assumption that the channel will approximately be the same during this period. As pointed out earlier, this condition can be easily satisfied. Again, all of this is done for each sub-carrier used to transmit STBC blocks of pilot /TPS data.

What has been described is merely illustrative of the application of the principles of the invention. Other arrangements and methods can be implemented by those skilled in

the art without departing from the spirit and scope of the present invention.

We claim:

1. A channel quality measurement apparatus adapted to measure a quality of a channel over which has been transmitted a sequence of symbols produced by encoding and constellation mapping a source data element sequence, the apparatus comprising:

a symbol de-mapper, receiving as input a sequence of received symbols over the channel whose quality is to be measured, said symbol de-mapper being adapted to perform symbol de-mapping on said sequence of received symbols to produce a sequence of soft data element decisions;

a soft decoder, receiving as input the sequence of soft data element decisions produced by the symbol de-mapper, said soft decoder being adapted to decode the sequence of soft data element decisions to produce a decoded output sequence;

an encoder, receiving as input the decoded output sequence produced by the soft decoder, said encoder being adapted to re-encode the decoded output sequence with an identical code to a code used in encoding the source data element sequence to produce a re-encoded output sequence; and

a correlator, receiving as input the sequence of soft data element decisions produced by the de-mapper, and the re-encoded output sequence produced by the encoder, said correlator being adapted to produce a channel quality indicator output by determining a correlation between the sequence of soft data element decisions and the re-encoded output sequence.

2. A channel quality measurement apparatus according to claim 1 wherein the symbol de-mapper is adapted to perform QPSK symbol de-mapping.

3. A channel quality measurement apparatus according to claim 1 wherein the symbol de-mapper is adapted to perform Euclidean distance conditional LLR symbol de-mapping.

4. A method of measuring channel quality of a channel over which has been transmitted a sequence of symbols produced by encoding and constellation mapping a source data element sequence, the method comprising:

receiving a sequence of received symbols over the channel whose quality is to be measured;

symbol de-mapping said sequence of received symbols to produce a sequence of soft data element decisions;

decoding said sequence of soft data element decisions to produce a decoded output sequence;

re-encoding said decoded output sequence to produce a re-encoded output sequence using a code identical to a code used in encoding the source data element sequence; and

correlating said re-encoded output sequence, and said sequence of soft data element decisions to produce a channel quality indicator output.

5. A method of channel quality measurement according to claim 4 wherein the symbol de-mapping of said sequence of received symbols is QPSK symbol de-mapping.

6. A method of channel quality measurement according to claim 4 wherein the symbol de-mapping of said sequence of received symbols comprises Euclidean distance conditional LLR de-mapping.

7. A method of measuring OFDM channel quality of an OFDM channel over which has been transmitted a sequence of OFDM symbols, the OFDM symbols containing an encoded and

constellation mapped source data element sequence, the method comprising:

receiving a sequence of OFDM symbols over the OFDM channel whose quality is to be measured;

symbol de-mapping said sequence of received symbols to produce a sequence of soft data element decisions;

decoding said sequence of soft data element decisions to produce a decoded output sequence pertaining to the source data element sequence;

re-encoding said decoded output sequence to produce a re-encoded output sequence using a code identical to a code used in encoding the source data element sequence; and

correlating said re-encoded output sequence, and said sequence of soft data element decisions to produce a channel quality indicator output.

8. A method of OFDM channel quality measurement according to claim 7 wherein the symbol de-mapping of said sequence of received symbols is QPSK symbol de-mapping.

9. A method of OFDM channel quality measurement according to claim 7 wherein the symbol de-mapping of said sequence of received symbols comprises Euclidean distance conditional LLR de-mapping.

10. A method of OFDM channel quality measurement according to claim 7 wherein the decoding of said sequence of soft data element decisions to produce a decoded output sequence further comprises using a history of the soft data element decisions, and using information about encoding of the sequence of symbols transmitted over the channel.

11. A communication system comprising:

a transmitter adapted to transmit a sequence of symbols produced by encoding and constellation mapping a source data element sequence over a channel; and

a receiver comprising:

a) a symbol de-mapper, receiving as input a sequence of received symbols over the channel, said symbol de-mapper being adapted to perform symbol de-mapping on said sequence of received symbols to produce a sequence of soft data element decisions;

b) a soft decoder, receiving as input the sequence of soft data element decisions produced by the symbol de-mapper, said soft decoder being adapted to decode the sequence of soft data element decisions to produce a decoded output sequence;

c) an encoder, receiving as input the decoded output sequence produced by the soft decoder, said encoder being adapted to re-encode the decoded output sequence with an identical code to a code used in encoding the source data element sequence to produce a re-encoded output sequence; and

d) a correlator, receiving as input the sequence of soft data element decisions produced by the de-mapper, and the re-encoded output sequence produced by the encoder, said correlator being adapted to produce a channel quality indicator output by determining a correlation between the sequence of soft data element decisions and the re-encoded output sequence,

wherein the receiver is adapted to feed the channel quality indicator back to the transmitter, and wherein the transmitter is adapted to use said channel quality indicator to determine and apply an appropriate coding rate and modulation to the source data element sequence.

12. A communication system according to claim 11 wherein the symbol de-mapper is adapted to perform QPSK symbol de-mapping.

13. A communication system according to claim 11 wherein the symbol de-mapper is adapted to perform Euclidean distance conditional LLR symbol de-mapping.

14. A method of adaptive modulation and coding comprising:

transmitting over a channel a sequence of symbols produced by encoding and constellation mapping a source data element sequence;

receiving a sequence of received symbols over the channel;

symbol de-mapping said sequence of received symbols to produce a sequence of soft data element decisions;

decoding said sequence of soft data element decisions to produce a decoded output sequence;

re-encoding said decoded output sequence to produce a re-encoded output sequence using a code identical to a code used in encoding the source data element sequence;

correlating said re-encoded output sequence, and said sequence of soft data element decisions to produce a channel quality indicator output;

transmitting the channel quality indicator; and

using said channel quality indicator to determine and apply an appropriate coding rate and modulation to the source data element sequence.

15. A method of adaptive modulation and coding according to claim 14 wherein the symbol de-mapping of said sequence of received symbols is QPSK symbol de-mapping.

16. A method of adaptive modulation and coding according to claim 14 wherein the symbol de-mapping of said sequence of received symbols comprises Euclidean distance conditional LLR de-mapping.

17. A method of determining a channel quality comprising correlating a soft data element decision sequence with a second data element sequence, the second data element sequence being produced by decoding the soft data element decision sequence to produce a decoded sequence and then re-encoding the decoded sequence.

18. A method comprising:

applying forward error coding to a signalling message to generate a coded fast signalling message;

MPSK mapping the coded signalling message to produce an MPSK mapped coded signalling message;

mapping the MPSK mapped coded signalling message onto a plurality of sub-carriers within an OFDM frame comprising a plurality of OFDM symbols;

encoding symbols of the MPSK mapped coded signalling message using Differential Space-Time Block Coding (D-STBC) in a time direction to generate encoded symbols; and

transmitting the encoded symbols on a plurality of transmit antennas, with the encoded symbols being transmitted at an increased power level relative to other symbols within the OFDM frame as a function of channel conditions.

19. A method according to claim 18 wherein the encoded symbols are transmitted in a scattered pattern.

20. A method according to claim 18 wherein transmitting the encoded symbols on a plurality of antennas comprises:

on a selected sub-carrier, each antenna transmitting a respective plurality N of encoded symbols over N consecutive OFDM symbols, where N is the number of antennas used to transmit, for a total of NxN transmitted encoded symbols, the NxN symbols being obtained from D-STBC encoding L symbols of the MPSK mapped coded signalling stream, where L,N determine an STBC code rate.

21. A method according to claim 20 further comprising:

transmitting a set of pilot sub-carriers in at least one OFDM symbol;

using the pilot sub-carriers as a reference for a first set of D-STBC encoded symbols transmitted during subsequent OFDM symbols.

22. A method according to claim 21 wherein transmitting a set of pilot sub-carriers in at least one OFDM frame comprises:

transmitting a plurality of pilots on each antenna on a respective disjoint plurality of sub-carriers.

23. A method according to claim 22 wherein each disjoint plurality of sub-carriers comprises a set of sub-carriers each separated by N-1 sub-carriers, where N is the number of antennas.

24. A method according to claim 22 wherein pilot sub-carriers are transmitted for a number of consecutive OFDM frames equal to the number of transmit antennas.

25. A method according to claim 18 wherein the signalling message contains an identification of one or more receivers who are to receive data during a current TPS frame.

26 An OFDM transmitter adapted to implement a method according to claim 18.

27. An OFDM transmitter adapted to implement a method according to claim 20.

28. A receiving method for an OFDM receiver comprising:

receiving at at least one antenna an OFDM signal containing received D-STBC coded MPSK mapped coded signalling message symbols;

recovering received signalling message symbols from the OFDM signal(s);

determining from the signalling message symbols whether a current OFDM transmission contains data to be recovered by the receiver;

upon determining the current OFDM transmission contains data to be recovered by the receiver:

a) re-encoding, MPSK mapping and D-STBC coding the received coded signalling message symbols to produce re-encoded D-STBC coded MPSK mapped coded signalling message symbols;

b) determining a channel estimate by comparing the received D-STBC coded mapped coded signalling message symbols with the re-encoded D-STBC coded MPSK mapped coded signalling message symbols.

29. A method according to claim 28 wherein a channel estimate is determined for each location (in time,frequency) in the OFDM signal containing D-STBC coded MPSK mapped coded

signalling message symbols, the method further comprising interpolating to get a channel estimate for remaining each location (in time, frequency) in the OFDM signal.

30. A method according to claim 29 further comprising: receiving pilot symbols which are not D-STBC encoded which are used as a reference for a first D-STBC block of D-STBC coded MPSK mapped coded signalling message symbols.

31. A method according to claim 28 further comprising:

extracting the signalling message.

32. An OFDM receiver adapted to implement the method of claim 28.

33. An article of manufacture comprising a computer-readable storage medium, the computer-readable storage medium including instructions for implementing the method of claim 1.

34. An article of manufacture comprising a computer-readable storage medium, the computer-readable storage medium including instructions for implementing the method of claim 18.

35. An article of manufacture comprising a computer-readable storage medium, the computer-readable storage medium including instructions for implementing the method of claim 28.

36. A method of generating pilot symbols from an Orthogonal Frequency Division Multiplexing (OFDM) frame received at an OFDM receiver, the OFDM frame containing an encoded fast signalling message in the form of encoded symbols within the OFDM frame, the method comprising the steps of:

processing the encoded symbols based in a scattered pilot pattern to recover the encoded fast signalling message;

re-encoding the fast signalling message so as to generate pilot symbols in the scattered pattern;

recovering a channel response for the encoded symbols using decision feedback.

37. The method of claim 36 comprising the further step of applying a fast algorithm to compute a Discrete Fourier Transform based on the scattered pattern to extract the combined pilot symbols and fast signalling message and only preceding to recover channel response if the fast signalling message indicates a current transmission contains content for the OFDM receiver.

38. The method of claim 35 wherein processing the encoded symbols comprises:

differentially decoding the encoded symbols using Differential Space-Time Block Coding (D-STBC) decoding to recover the encoded fast signalling message;

applying Forward Error Correction decoding to the encoded fast signalling message to recover a fast signalling message;

analyzing the fast signalling message to determine whether it includes a desired user identification;

if the fast signalling message includes the desired user identification, re-encoding the fast signalling message using Forward Error Correction coding to generate the encoded fast signalling message, and re-encoding the encoded fast signalling message using D-STBC.

39. A transmitter adapted to combine pilot and transmission parameter signalling on a single overhead channel within an OFDM signal.

40. A transmitter according to claim 39 wherein a set of transmission parameter signalling symbols are transmitted on the overhead channel with strong encoding such that at a receiver, they can be decoded accurately, re-encoded, and the re-encoded symbols treated as known pilot symbols which can then be used for channel estimation.

41. A receiver adapted to process the combined single overhead channel produced by the transmitter of claim 40, the receiver being adapted to:

decode a received signal containing the encoded transmission parameter signalling symbols as modified by a channel, re-encode the decoded symbols to produce known pilot symbols, compare received symbols with the known pilot symbols to produce a channel estimate.

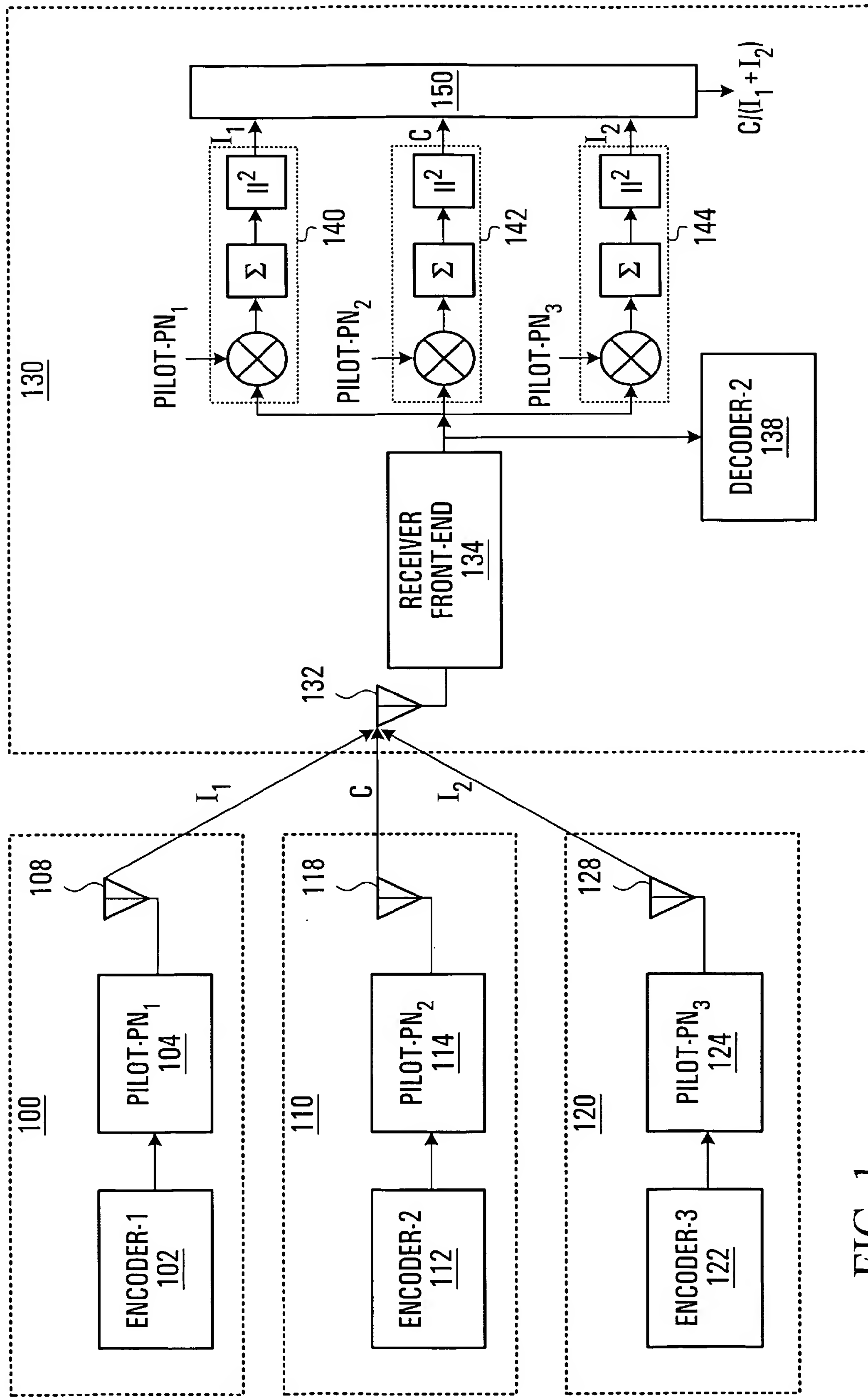


FIG. 1
(PRIOR ART)

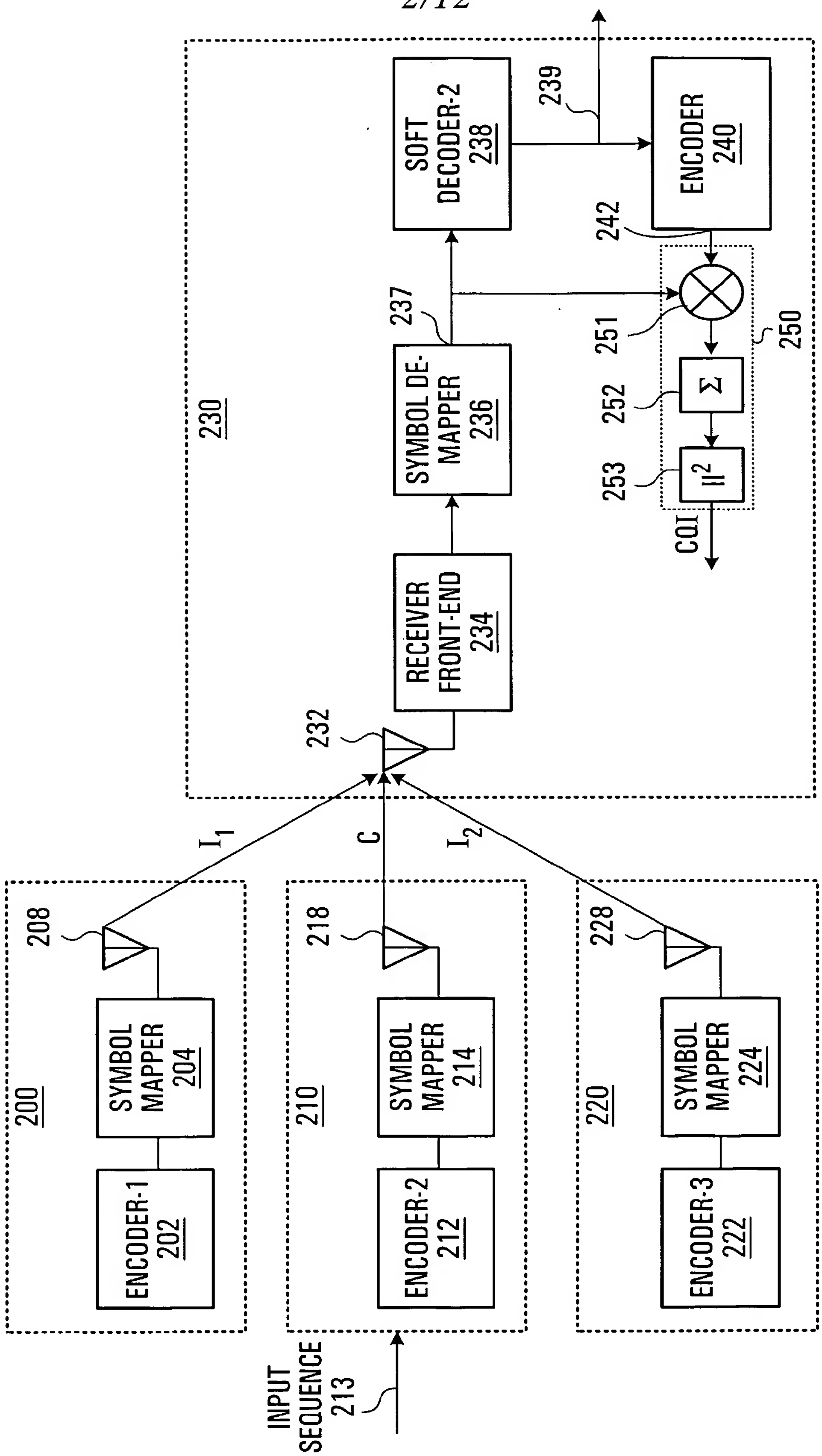

200
ENCODER-1 202
SYMBOL MAPPER 204
208
210
ENCODER-2 212
SYMBOL MAPPER 214
218
INPUT SEQUENCE 213
220
ENCODER-3 222
SYMBOL MAPPER 224
228
I1
C
I2
230
232
RECEIVER FRONT-END 234
SYMBOL DE-MAPPER 236
237
SOFT DECODER-2 238
239
ENCODER 240
242
251
250
Σ
252
253
CQI

FIG. 2

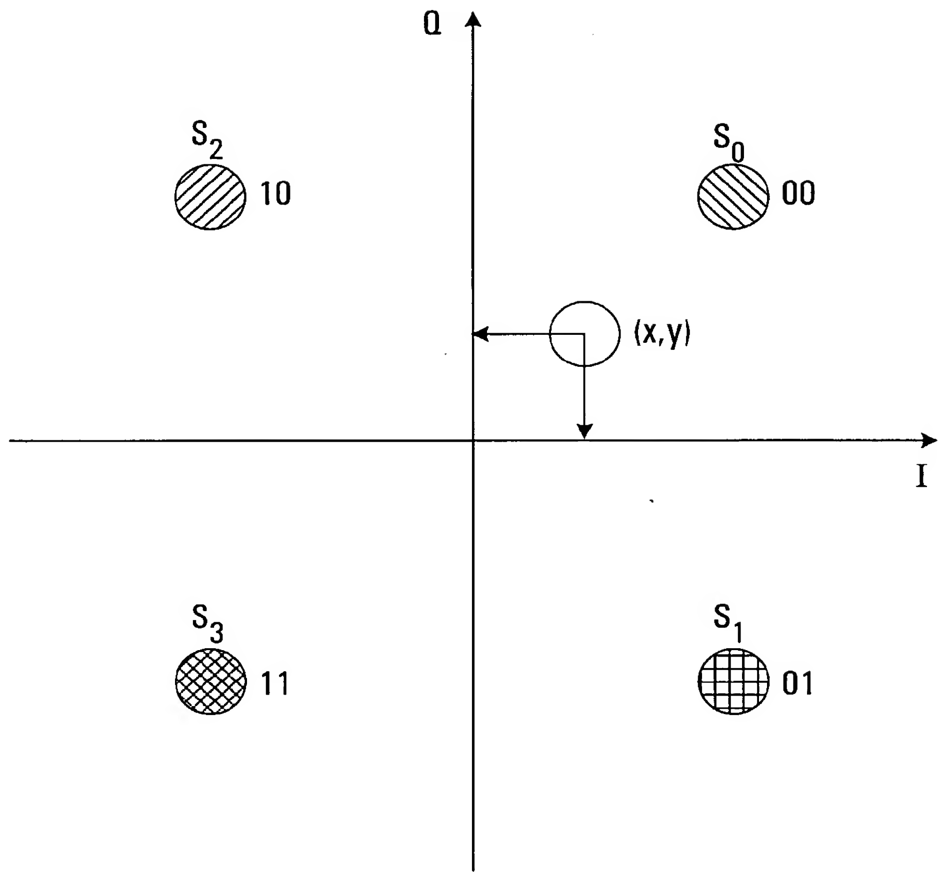


FIG. 3

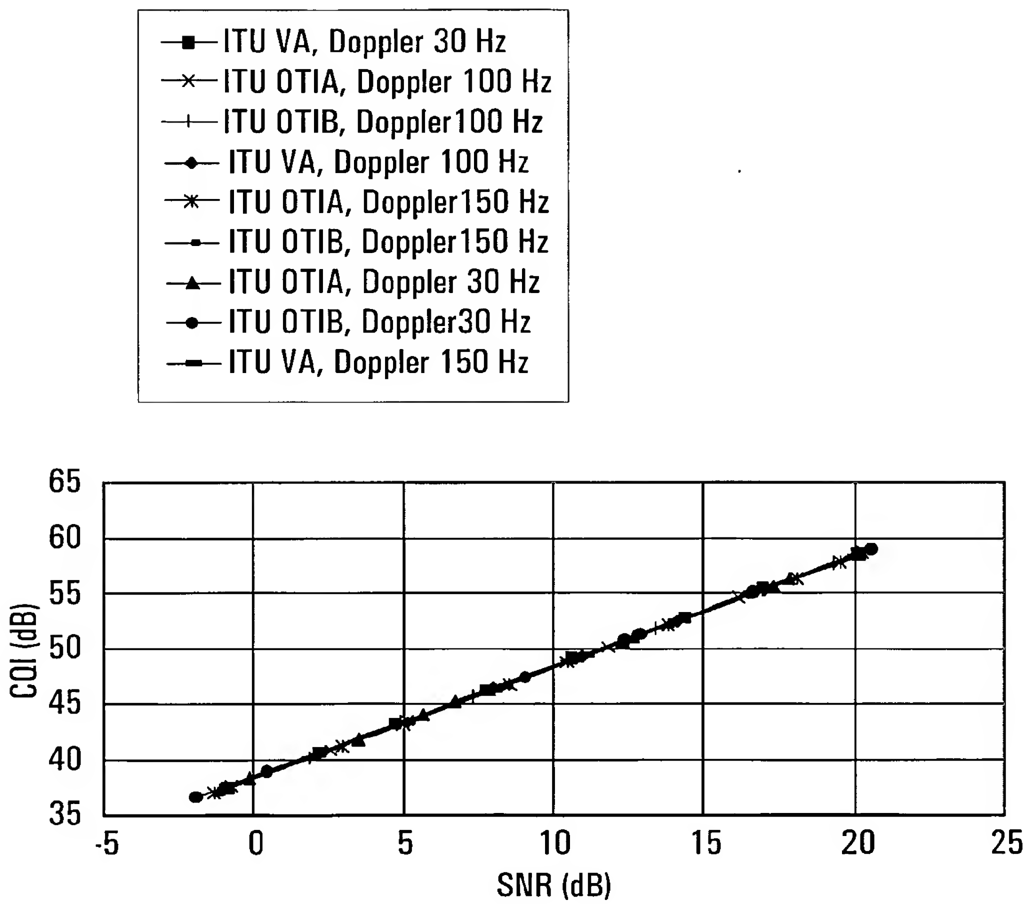

ITU VA, Doppler 30 Hz
ITU OTIA, Doppler 100 Hz
ITU OTIB, Doppler100 Hz
ITU VA, Doppler 100 Hz
ITU OTIA, Doppler150 Hz
ITU OTIB, Doppler150 Hz
ITU OTIA, Doppler 30 Hz
ITU OTIB, Doppler30 Hz
ITU VA, Doppler 150 Hz
65
60
55
50
45
40
35
CQI (dB)
-5
0
5
10
15
20
25
SNR (dB)

FIG. 4

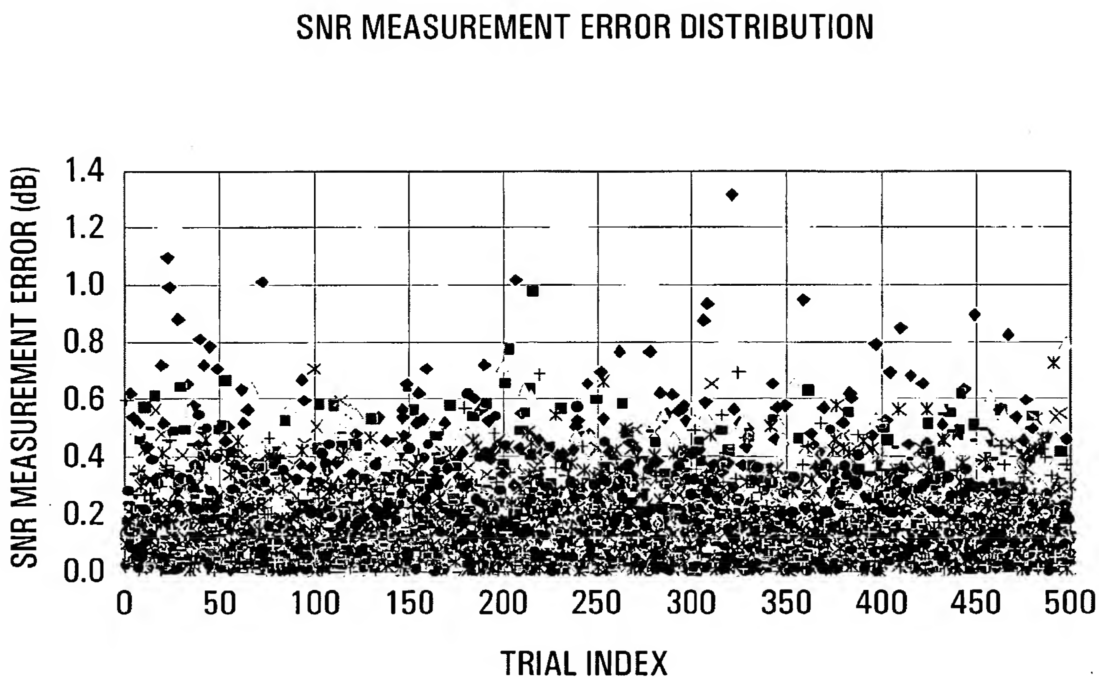

SNR MEASUREMENT ERROR DISTRIBUTION
SNR MEASUREMENT ERROR (dB)
1.4
1.2
1.0
0.8
0.6
0.4
0.2
0.0
0
50
100
150
200
250
300
350
400
450
500
TRIAL INDEX

FIG. 5

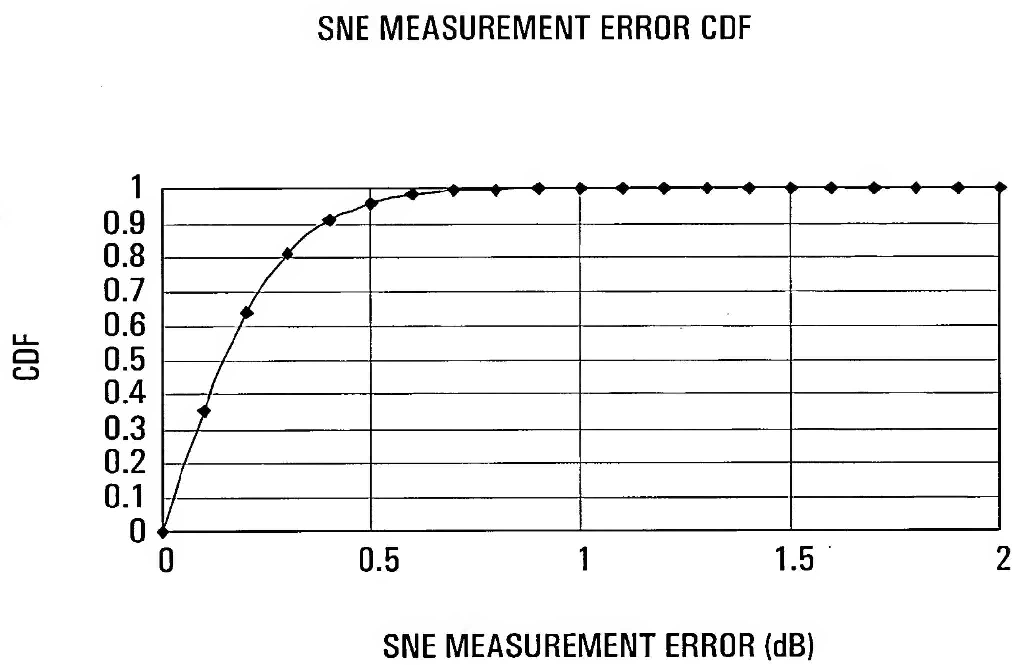

SNE MEASUREMENT ERROR CDF
CDF
1
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
0
0.5
1
1.5
2
SNE MEASUREMENT ERROR (dB)

FIG. 6

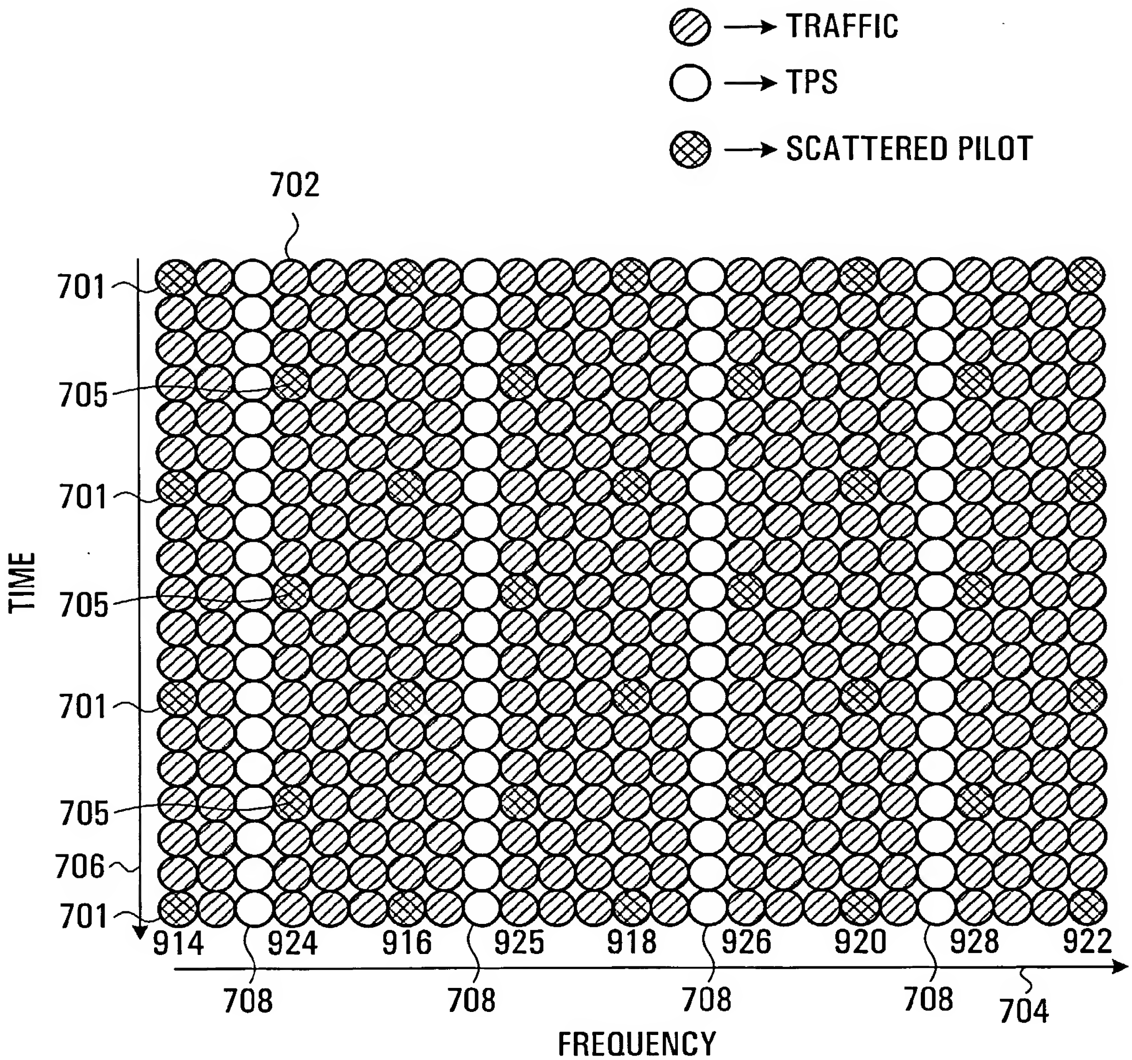


FIG. 7
(PRIOR ART)

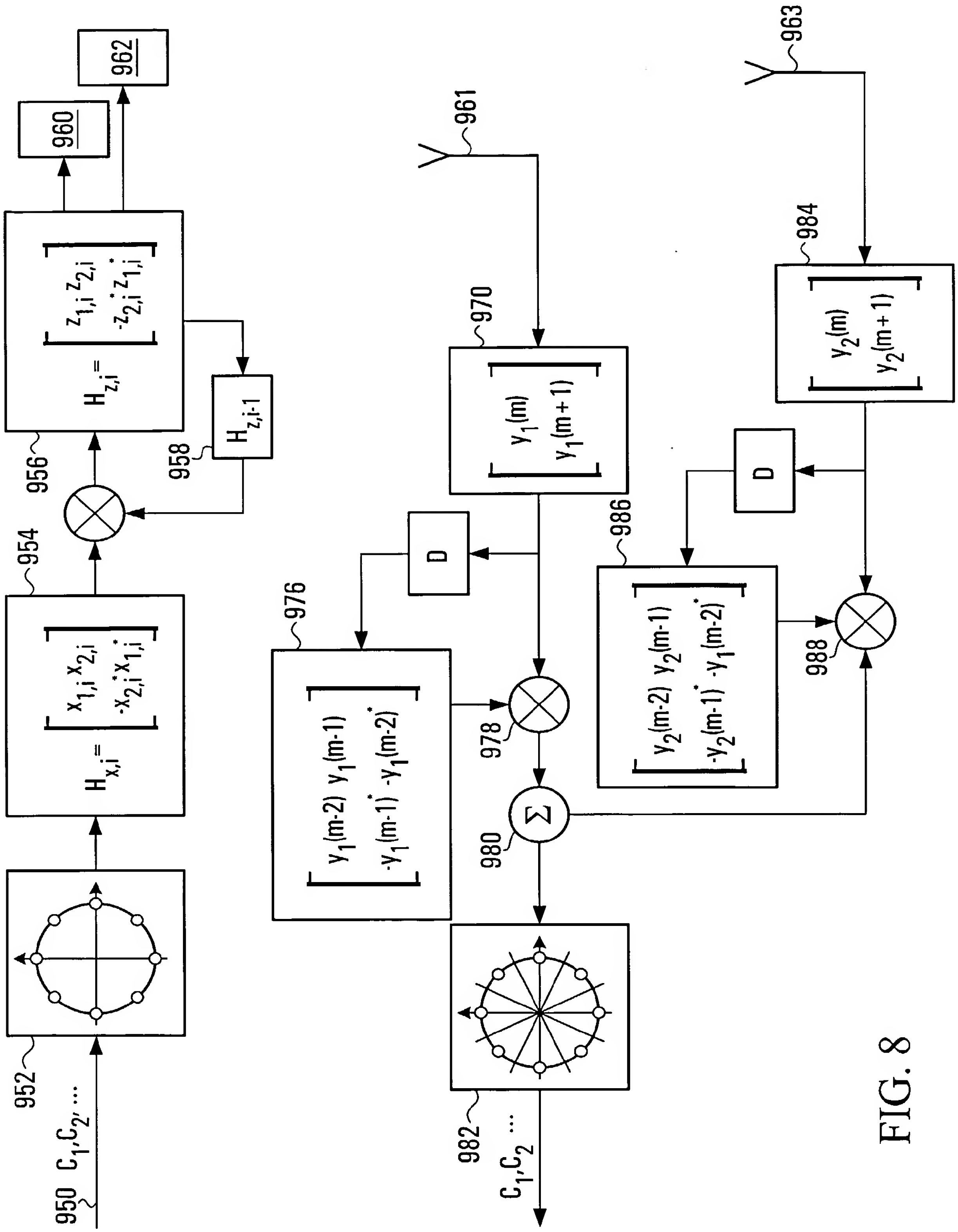

950
$C_1, C_2, \ldots$
952
954
$H_{x,i} = \begin{bmatrix} x_{1,i} & x_{2,i} \\ -x_{2,i}^* & x_{1,i}^* \end{bmatrix}$
956
$H_{z,i} = \begin{bmatrix} z_{1,i} & z_{2,i} \\ -z_{2,i}^* & z_{1,i}^* \end{bmatrix}$
958
$H_{z,i-1}$
960
962
961
963
970
$\begin{bmatrix} y_1(m) \\ y_1(m+1) \end{bmatrix}$
984
$\begin{bmatrix} y_2(m) \\ y_2(m+1) \end{bmatrix}$
D
D
976
$\begin{bmatrix} y_1(m-2) & y_1(m-1) \\ -y_1(m-1)^* & -y_1(m-2)^* \end{bmatrix}$
986
$\begin{bmatrix} y_2(m-2) & y_2(m-1) \\ -y_2(m-1)^* & -y_1(m-2)^* \end{bmatrix}$
978
988
980
Σ
982
$C_1, C_2 \ldots$

FIG. 8

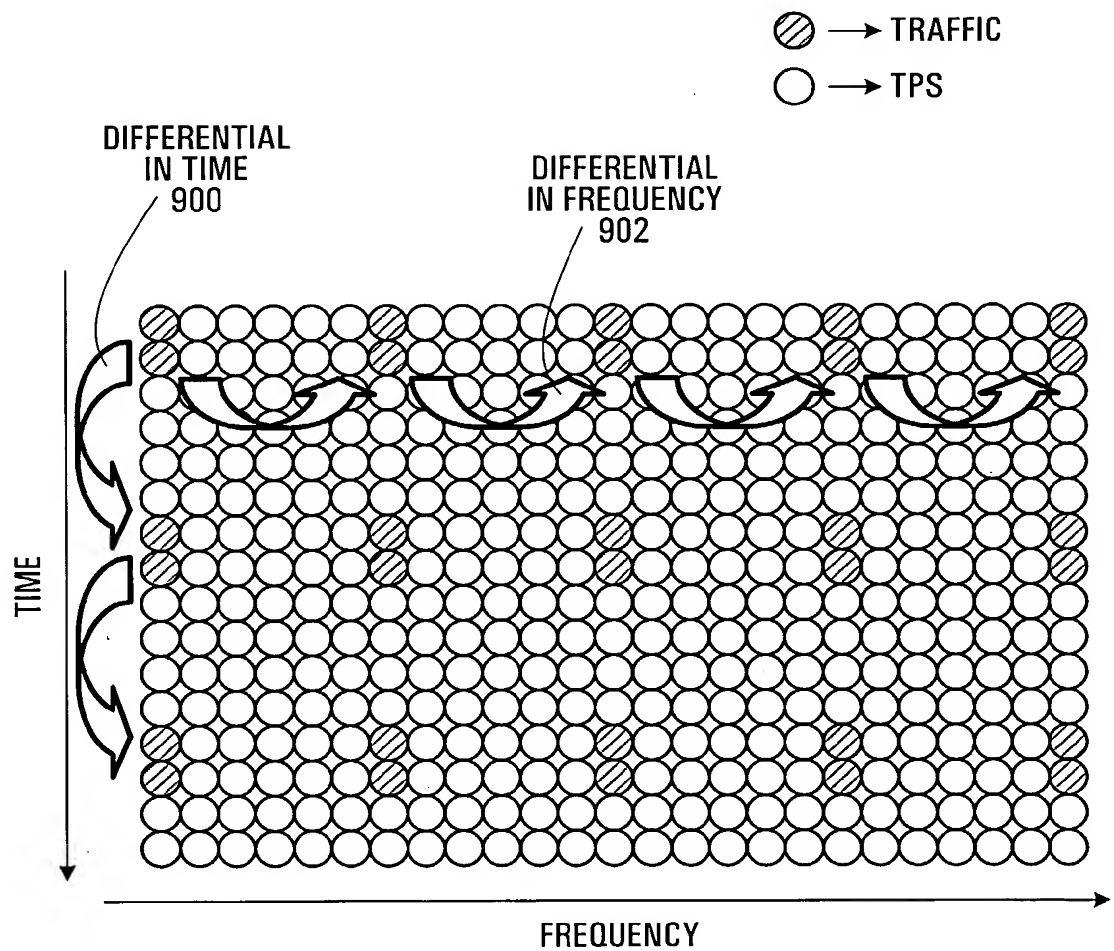
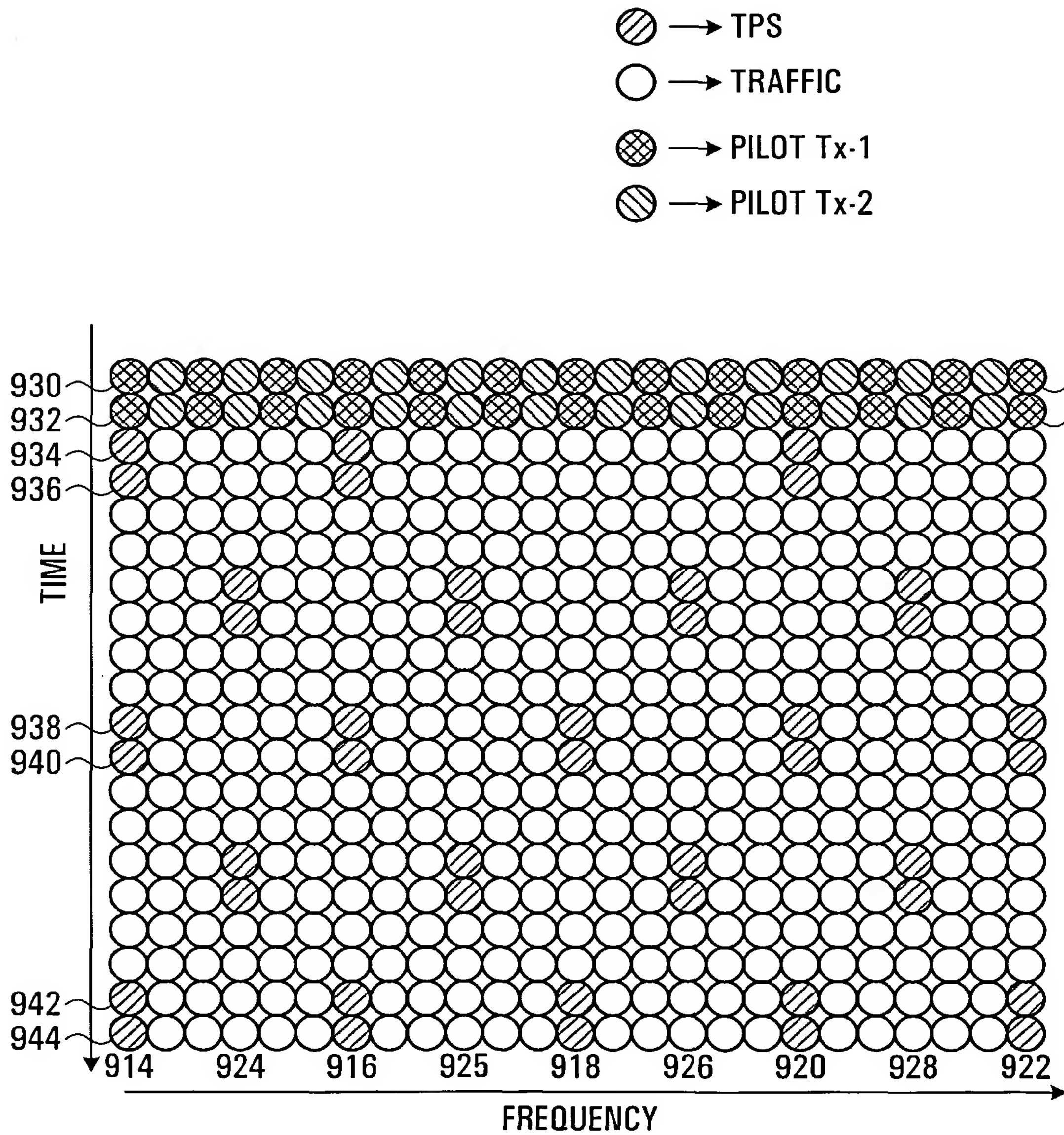

TRAFFIC
TPS
DIFFERENTIAL IN TIME 900
DIFFERENTIAL IN FREQUENCY 902
TIME
FREQUENCY

FIG. 9



TPS
TRAFFIC
PILOT Tx-1
PILOT Tx-2
930
932
934
936
TIME
938
940
942
944
910
912
914
924
916
925
918
926
920
928
922
FREQUENCY

FIG. 10

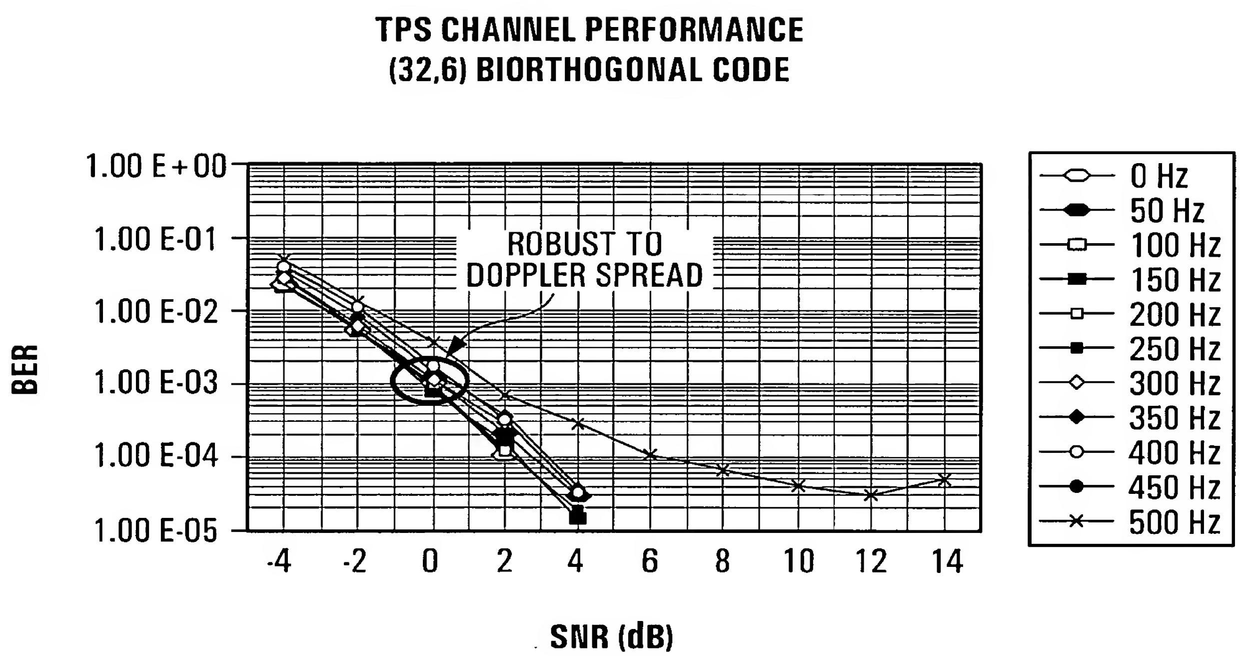

TPS CHANNEL PERFORMANCE
(32,6) BIORTHOGONAL CODE
BER
1.00 E+00
1.00 E-01
1.00 E-02
1.00 E-03
1.00 E-04
1.00 E-05
ROBUST TO DOPPLER SPREAD
-4
-2
0
2
4
6
8
10
12
14
SNR (dB)
0 Hz
50 Hz
100 Hz
150 Hz
200 Hz
250 Hz
300 Hz
350 Hz
400 Hz
450 Hz
500 Hz

FIG. 11

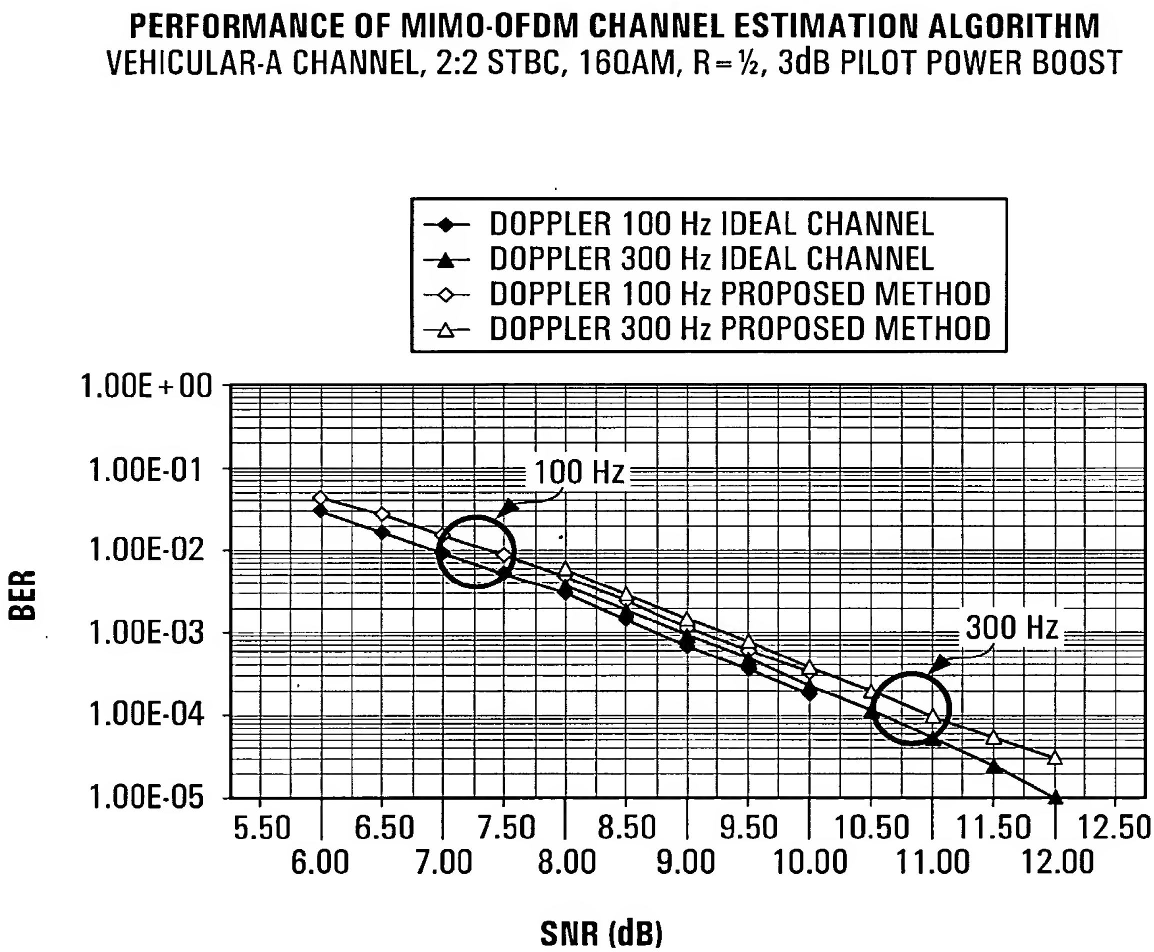

PERFORMANCE OF MIMO-OFDM CHANNEL ESTIMATION ALGORITHM
VEHICULAR-A CHANNEL, 2:2 STBC, 16QAM, R=½, 3dB PILOT POWER BOOST
DOPPLER 100 Hz IDEAL CHANNEL
DOPPLER 300 Hz IDEAL CHANNEL
DOPPLER 100 Hz PROPOSED METHOD
DOPPLER 300 Hz PROPOSED METHOD
100 Hz
300 Hz
BER
1.00E+00
1.00E-01
1.00E-02
1.00E-03
1.00E-04
1.00E-05
5.50
6.00
6.50
7.00
7.50
8.00
8.50
9.00
9.50
10.00
10.50
11.00
11.50
12.00
12.50
SNR (dB)

FIG. 12

ORTHOGONAL FREQUENCY-DIVISION MULTIPLEX TRANSMISSION METHOD

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Inventor(s): NAKAMURA TAKAHARU [JP] +
Applicant(s): FUJITSU LTD [JP]; NAKAMURA TAKAHARU [JP] +
Classification:
- international: ***H04B7/005; H04J11/00; H04L27/26; H04L5/02;*** **H04B7/005; H04J11/00; H04L27/26; H04L5/02;** (IPC1-7): H04J11/00
- European: H04J11/00; H04L27/26M1G; H04L5/00A3A; H04W52/26R; H04W52/28
Application number: WO2001JP10357 20011128
Priority number(s): WO2001JP10357 20011128

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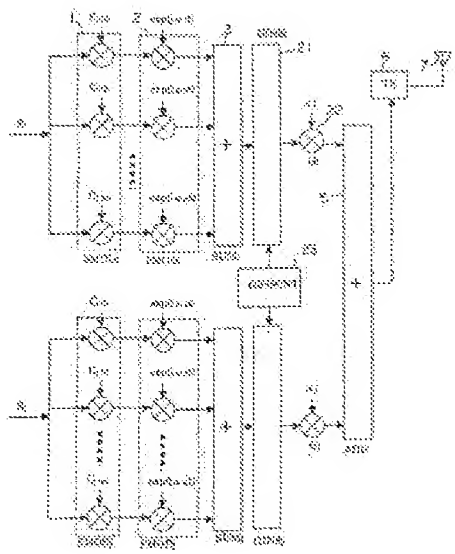
EP1450505 (B1)
US2004213145 (A1)
US2007183310 (A1)
more >>

Cited documents:
EP1035693 (A2)
JP2000165342 (A)
EP1014639 (A2)
JP2001111519 (A)
JP11196062 (A)

Abstract of **WO 03047140 (A1)**

A spreading modulator 1 spreads the spectrum of a signal series. A subcarrier modulator 2 modulates the frequencies of a plurality of subcarriers having frequencies different from one another by using the output of the spreading modulator 1. An adder 3 combines the subcarriers modulated. A guard section control unit 23 determines the length of a guard section in accordance with the maximum transmission delay difference of a line between a transmission device and a reception device. A guard section inserter 21 is controlled by the guard section control unit 23 to insert the guard section into the signal series for every symbol period. A gain adjustor 22 multiplies the transmission signal by a gain coefficient corresponding to the guard section inserted.A spreading modulator (1) spreads the spectrum of a signal series. A subcarrier modulator (2) modulates the frequencies of a plurality of subcarriers having frequencies different from one another by using the output of the spreading modulator (1). An adder (3) combines the subcarriers modulated. A guard section control unit (23) determines the length of a guard section in accordance with the maximum transmission delay difference of a line between a transmission device and a reception device. A guard section inserter (21) is controlled by the guard section control unit (23) to insert the guard section into the signal series for every symbol period. A gain adjustor (22) multiplies the transmission signal by a gain coefficient corresponding to the guard section inserted.



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国際事務局

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(26) 国際公開の言語: 日本語

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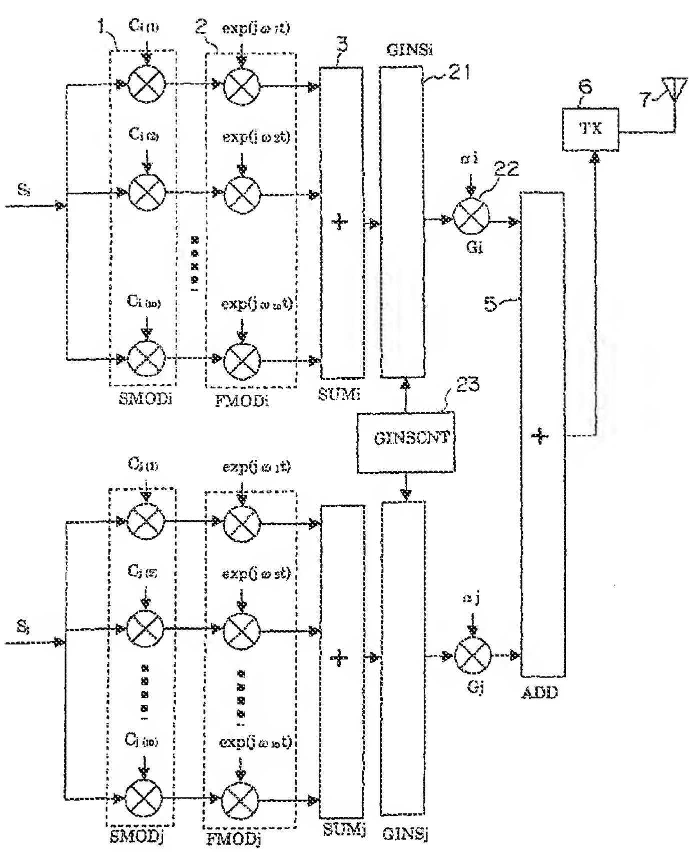
(74) 代理人: 大菅義之(OSUGA, Yoshiyuki); 〒102-0084 東京都千代田区二番町8番地20 二番町ビル3階 Tokyo (JP).

(81) 指定国 (国内): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO,

[続葉有]

(54) Title: ORTHOGONAL FREQUENCY-DIVISION MULTIPLEX TRANSMISSION METHOD

(54) 発明の名称: 直交周波数分割多重伝送方法



(57) Abstract: A spreading modulator (1) spreads the spectrum of a signal series. A subcarrier modulator (2) modulates the frequencies of a plurality of subcarriers having frequencies different from one another by using the output of the spreading modulator (1). An adder (3) combines the subcarriers modulated. A guard section control unit (23) determines the length of a guard section in accordance with the maximum transmission delay difference of a line between a transmission device and a reception device. A guard section inserter (21) is controlled by the guard section control unit (23) to insert the guard section into the signal series for every symbol period. A gain adjustor (22) multiplies the transmission signal by a gain coefficient corresponding to the guard section inserted.

WO 03/047140 A1

[続葉有]

NZ, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, US, UZ, VN, YU, ZA, ZW.

(84) 指定国 (広域): ARIPO 特許 (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), ユーラシア特許 (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), ヨーロッパ特許 (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI 特許 (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

添付公開書類:
— 国際調査報告書

2文字コード及び他の略語については、定期発行される各PCTガゼットの巻頭に掲載されている「コードと略語のガイダンスノート」を参照。

(57) 要約:

拡散変調器(1)は、信号系列を拡散する。副搬送波変調器(2)は、拡散変調器(1)の出力を用いて互いに周波数の異なる複数の副搬送波を周波数変調する。加算器(3)は、変調された各副搬送波を合成する。ガード区間制御部(23)は、送信装置と受信装置との間の回線の最大伝送遅延差に応じてガード区間の長さを決定する。ガード区間挿入器(21)は、シンボル周期ごとに、ガード区間制御部(23)の制御に従って信号系列にガード区間を挿入する。利得調整器(22)は、挿入されたガード区間に対応する利得係数を送信信号に乗算する。

明細書

直交周波数分割多重伝送方法

技術分野

本発明は、直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式、並びにそのための送信装置（変調装置）及び受信装置（復調装置）に係わり、特に、セルラ電話システムまたは移動体通信システムにおける基地局と移動局との間の通信を実現する装置および方法に係わる。

背景技術

従来より、地上系デジタルテレビ等において、直交周波数分割多重（以下、ＯＦＤＭ：Orthogonal Frequency Division Multiplex）伝送方式が適用されている。ＯＦＤＭ伝送方式では、データは、互いに周波数の異なる複数の副搬送波を利用して伝送される。具体的には、この方式では、互いに直交する多数の副搬送波を送信データで変調し、それらの副搬送波が周波数多重されて伝送される。そして、ＯＦＤＭ伝送方式によれば、高速データの伝送を行う場合においても、各副搬送波ごとの伝送レートを低くできるので、すなわち各副搬送波ごとのシンボル周期を長くできるので、マルチパス干渉の影響が軽減される。なお、ＯＦＤＭ伝送方式につては、例えば、"Overview of Multicarrier CDMA" (Hara et al., IEEE Communication Magazine, Dec. 1997, pp126-133)、あるいは、"WIDEBAND WIRELESS DIGITAL COMMUNICATIONS", A.F.Molisch Prentice Hall PTR, 2001, ISBN:0-13-022333-6)に記載されている。

図１は、ＯＦＤＭ伝送システムにおいて使用される既存の送信装置の構成図である。ここでは、この送信装置は、信号系列Ｓｉ および信号系列Ｓｊ を多重化

して出力するものとする。なお、信号系列Ｓi および信号系列Ｓj のシンボル周期は「T」であるものとする。また、信号系列Ｓi および信号系列Ｓj は、例えば、互いに異なる移動機へ送信すべき信号であってもよい。あるいは、信号系列Ｓi 内に複数の移動機へ送信すべきデータが時間多重されていてもよい。

信号系列Ｓi の各シンボル情報は、それぞれ拡散変調器１が備えるｍ個の入力端子に並列に入力される。すなわち、拡散変調器１の各入力端子には、シンボル周期Ｔとごに、同一のシンボル情報が並列に入力される。そして、拡散変調器１は、入力されたシンボル情報を信号系列Ｓi に対して予め割り当てられている拡散符号Ｃi を用いて変調し、その結果として得られるｍビットの拡散信号を出力する。なお、拡散符号Ｃi は、「Ｃi(1)」～「Ｃi(m)」から構成されており、直交符号列の中の１つの要素であるものとする。

副搬送波変調器２は、互いに異なる角周波数ω1～ωm を持ったｍ個の副搬送波を生成する。ここで、ω1, ω2, ω3, . . . ωm の角周波数間隔Δωは、シンボル周期Ｔの逆数により定義される一定の値であり、下記の式により表される。

$\Delta\omega = 2\pi\Delta f = 2\pi / T$

また、副搬送波変調器２は、拡散変調器１から出力される拡散信号を用いてｍ個の副搬送波を変調する。具体的には、例えば、角周波数ω1 を持った副搬送波は、「Ｃi(1)」が乗算されたシンボル情報により変調され、角周波数ωm を持った副搬送波は、「Ｃi(m)」が乗算されたシンボル情報により変調される。そして、各副搬送波は、加算器３により合成される。

ガード区間挿入器４は、図２に示すように、シンボル毎に、加算器３から出力される合成信号に対して、予め固定的に決められているガード区間（Guard Interval）を挿入する。ここで、このガード区間は、無線伝送路のマルチパスによる影響を排除するために挿入される。なお、図２では、副搬送波ごとにガード区間が挿入された状態が描かれているが、実際には、これらの副搬送波は合

成されている。

加算器５は、上述のようにして得られる信号系列Ｓｉに対応する合成信号、および同様の処理により得られる信号系列Ｓｊに対応する合成信号を加算する。ここで、信号系列Ｓｉ に対応する合成信号および信号系列Ｓｊ に対応する合成信号には、それぞれガード区間が挿入されている。そして、加算器５の出力は、送信機６により所定の高周波信号に変換された後、アンテナ７を介して送信される。

図３は、ＯＦＤＭ伝送システムにおいて使用される既存の受信装置の構成図である。ここでは、この受信装置は、図１に示す送信装置により送信された無線信号から信号系列Ｓｉを受信するものとする。なお、図３では、信号を受信するために必要な周波数同期機能、およびタイミング同期機能などは省略されている。

アンテナ１１により受信された信号は、受信機１２によりベースバンド信号Ｓrx に変換された後、副搬送波復調器１３によりｍ個の受信信号列に変換される。続いて、ガード区間削除器１４は、各受信信号列からそれぞれガード区間を削除する。また、拡散復調器１５は、各受信信号系列を逆拡散するために、送信装置において使用された拡散符号と同じ拡散符号Ｃｉ を各受信信号列にそれぞれ乗算する。そして、拡散復調器１５から出力される各信号を加算器１６を用いて加算することにより、信号系列Ｓｉが再生される。

上記構成の送信装置および受信装置の間では、信号系列Ｓｉは、図２に示すように、複数の副搬送波ｆ１～ｆｍ を利用して伝送される。ここで、信号系列Ｓｉは、「＋１」または「－１」の値を有するシンボル情報から構成されている。即ち、信号系列Ｓｉは、シンボル周期Ｔで「＋１」または「－１」に変化する。また、各副搬送波ｆ１～ｆｍ を利用して伝送される信号は、それぞれ拡散符号Ｃｉ（Ci(1), Ci(2), . . . Ci(m)）により拡散変調されている。なお、図２において、

「＊」が付されているビットは、信号系列Ｓｉが「－１」であることから拡散変調出力が反転（共役）出力になっていることを示している。

伝送される信号には、上述したように、シンボル毎にガード区間が挿入されている。図２に示す例では、シンボル周期Ｔに対してガード区間Ｔｇが挿入されている。したがって、受信装置では、各副搬送波ごとにそれぞれガード区間Ｔｇを除去することにより得られる区間（区間Ｔs）について逆拡散／復調処理が行われる。これにより、受信装置においてマルチパス干渉（遅延波により生ずる干渉）が除去される。

ところで、ガード区間Ｔｇは、マルチパス干渉を除去するために挿入されるので、その長さは、伝送路の最大伝送遅延差よりも長く設定される必要がある。ここで、「最大伝送遅延差」とは、送信装置から受信装置へ複数のパスを介して信号が伝送されるときの最小伝搬時間と最大伝搬時間との差を意味する。例えば、図４において、パス１を介して伝送された信号が最も早く受信装置に到着し、パス３を介して伝送された信号が最も遅く受信装置に到着したとすると、最大伝送遅延差は、パス３による伝搬時間とパス１による伝搬時間との差により表される。

ところが、セルラ通信システムでは、通常、１つの基地局からサービスエリア内の複数の移動機に対して無線信号が送信される。そして、基地局から移動機へ伝送される信号の最大伝送遅延差は、一般に、それらの間の距離が離れるほど大きくなる傾向にある。ここで、サービスエリア内のすべての移動機においてマルチパス干渉を除去しようとすると、基地局から最も遠く離れた位置にいる移動機においてマルチパス干渉を除去できるようにしなければならない。したがって、サービスエリア内のすべての移動機においてマルチパス干渉を除去しようとすると、ガード区間Ｔｇは、基地局から最も遠く離れた位置にいる移動機に信号が伝送されたときの最大伝送遅延差よりも大きくする必要がある。

例えば、図５に示す例では、ガード区間Ｔｇは、基地局から移動機ＭＳ３に信号が伝送されたときの最大伝送遅延差よりも大きくする必要がある。

しかし、このようにしてガード区間の差を決定すると、基地局の近くに位置している移動機（図５では、移動機ＭＳ１）に信号を送信する場合には、ガード区間が必要以上に長くなりすぎる。ここで、ガード区間の信号の電力は、受信装置において信号系列を再生する際に使用されない。このため、上述のようにしてガード区間が決定されると、移動機に信号を送信する際に、無駄な電力が必要となってしまう。この結果、通信システム全体の総伝送容量の低減をまねくことになる。

発明の開示

本発明は、直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式を利用した通信システムにおいて、信号の伝送効率を向上させることを目的とする。

本発明の通信システムは、直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する通信システムであって、上記送信装置は、信号系列を用いて複数の副搬送波を変調する変調手段と、上記変調手段の出力にガード区間を挿入する挿入手段と、上記ガード区間が挿入された変調信号を送信する送信手段を有し、上記受信手段は、上記送信装置から送信された変調信号について副搬送波ごとにガード区間の削除処理と復調処理を行い信号系列を再生する復調手段を有し、上記ガード区間の長さは、上記送信装置と上記受信装置との間の通信環境に基づいて決定される。

上記通信システムにおいては、送信装置と受信装置との間の通信環境に基づいてガード区間の長さが決定される。すなわち、ガード区間の長さを、送信装置と受信装置との間の通信環境に応じて必要最小限に短くできる。したがって、

通信効率が向上する。

上記構成において、上記送信装置が、上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有するようにしてもよい。この構成によれば、信号系列を送信する際の送信電力を必要最小限に抑えることができるので、信号間の干渉が低減する。

上記構成において、上記受信装置が、上記送信装置から当該受信装置へ信号が伝送されたときの通信品質をモニタするモニタ手段をさらに有し、上記ガード区間の長さが、予め決められた所定の通信品質が満たされるように決定されるようにしてもよい。この構成によれば、所望の通信品質を満たす範囲内で、必要最小限のガード区間を設定できる。

本発明の他の態様の通信システムは、直交周波数分割多重を利用して送信装置から第１の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、上記送信装置は、第１の受信装置へ伝送する第１の信号系列および第１の受信装置とは異なる他の受信装置へ伝送する第２の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、上記第１の信号系列の変調出力に第１のガード区間を挿入するとともに上記第２の信号系列の変調出力に第２のガード区間を挿入する挿入手段と、上記第１のガード区間と第２のガード区間がそれぞれ挿入された変調信号を送信する送信手段を有し、上記第１の受信装置は、上記第１のガード区間の削除処理と復調処理を行い第１の信号系列を再生する復調手段を有し、上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定されると共に、上記第２のガード区間の長さは、上記送信装置と上記他の受信装置との間の通信環境に基づいて決定される。この構成によれば、複数の信号系列を時間多重して送信する際に、各信号系列に対して個々に適切なガード区間を設定できる。

図面の簡単な説明

図１は、ＯＦＤＭ伝送システムにおいて使用される既存の送信装置の構成図である。

図２は、既存のＯＦＤＭ伝送システムにおける伝送信号の例である。

図３は、ＯＦＤＭ伝送システムにおいて使用される既存の受信装置の構成図である。

図４は、マルチパスを説明する図である。

図５は、複数の移動機を収容する基地局を示す図である。

図６は、本発明の実施形態の送信装置の構成図である。

図７は、本発明の実施形態の受信装置の構成図である。

図８および図９は、実施形態のＯＦＤＭ伝送システムにおける伝送信号の例である。

図１０は、ガード区間について説明するための図である。

図１１は、副搬送波変調器により実行される逆フーリエ変換を説明する図である。

図１２は、ガード区間を挿入する処理を説明する図である。

図１３は、ガード区間を挿入する処理を実現する構成の実施例である。

図１４は、受信波からガード区間を削除する処理を実現する構成の実施例である。

図１５は、第１の実施例の送信装置の構成図である。

図１６は、第１の実施例の受信装置の構成図である。

図１７は、第１の実施例の通信システムにおける伝送信号を模式的に示す図である。

図１８は、第２の実施例の送信装置の構成図である。

図１９は、第２の実施例の受信装置の構成図である。

図２０は、第２の実施例の通信システムにおける伝送信号を模式的に示す図である。

図２１は、第３の実施例の送信装置の構成図である。

図２２は、第３の実施例の受信装置の構成図である。

図２３は、図２２に示す遅延差検出部の一例の構成図である。

図２４は、遅延差検出部の動作を説明する図である。

図２５は、最大伝送遅延差を検出する実施例である。

図２６は、第４の実施例の送信装置の構成図である。

図２７は、第４の実施例の受信装置の構成図である。

図２８は、図２７に示す距離推定部の一例の構成図である。

図２９は、第５の実施例の送信装置の構成図である。

図３０は、第５の実施例の受信装置の構成図である。

図３１は、図３０に示すタイミング生成部の一例の構成図である。

図３２は、第６の実施例の送信装置の構成図である。

図３３は、第６の実施例の受信装置の構成図である。

図３４は、図３３に示すタイミング生成部の一例の構成図である。

図３５は、第７の実施例の送信装置の構成図である。

図３６は、第７の実施例の受信装置の構成図である。

図３７は、図３６に示す遅延差検出部の動作を示すフローチャートである。

図３８は、第８の実施例の送信装置の構成図である。

図３９は、第８の実施例の受信装置の構成図である。

図４０は、図３９に示す距離推定部の動作を示すフローチャートである。

発明を実施するための最良の形態

本発明の実施形態について図面を参照しながら説明する。以下では、セルラ

通信システムにおいて直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式が利用されるものとする。具体的には、例えば、図５に示す基地局と移動機との間の信号伝送のためにＯＦＤＭ－ＣＤＭが利用されるものとする。

図６は、本発明の実施形態の送信装置の構成図である。なお、この送信装置は、図５においては、例えば、基地局装置に相当する。また、この送信装置は、信号系列Ｓi および信号系列Ｓj を多重化して出力するものとする。ここで、信号系列Ｓi および信号系列Ｓj は、例えば、互いに異なる移動機へ送信すべき信号であってもよい。あるいは、信号系列Ｓi または信号系列Ｓj の中にそれぞれ複数の移動機へ送信すべきデータが時間多重されていてもよい。

この送信装置は、送信すべき信号系列毎に、拡散変調器（ＳＭＯＤ：Spread Modulator）１、副搬送波変調器（ＦＭＯＤ：Frequency Modulator）２、加算器（ＳＵＭ）３、ガード区間挿入器（ＧＩＮＳ：Guard Interval Insert Unit）２１、利得調整器（Ｇ）２２を備える。ここで、拡散変調器１、副搬送波変調器２、加算器３については、図１を参照しながら説明したものを使用することができる。すなわち、拡散変調器１は、ｍ個の入力端子を備えており、それらの入力端子には、シンボル周期Ｔごに、同一のシンボル情報が並列に入力される。そして、拡散変調器１は、入力されたシンボル情報を信号系列Ｓiに対して予め割り当てられている拡散符号Ｃiを用いて変調し、その結果として得られるｍビットの拡散信号を出力する。なお、拡散符号Ｃi は、「Ｃi(1)」～「Ｃi(m)」から構成されており、直交符号列の中の１つの要素であるものとする。

副搬送波変調器２は、互いに異なる角周波数ω１～ωｍを持ったｍ個の副搬送波を生成する。ここで、ω１, ω２, ω３, . . . ωｍの角周波数間隔Δωは、シンボル周期Ｔの逆数により定義される一定の値であり、下記の式により表される。

$$\Delta\omega = 2\pi\Delta f = 2\pi / T$$

また、副搬送波変調器２は、拡散変調器１から出力される拡散信号を用いて

m個の副搬送波を変調する。具体的には、例えば、角周波数ω1を持った副搬送波は、「Ci(1)」が乗算されたシンボル情報により変調され、角周波数ωm を持った副搬送波は、「Ci(m)」が乗算されたシンボル情報により変調される。なお、副搬送波変調器２の処理は、例えば、逆フーリエ変換演算により実現される。そして、副搬送波変調器２から出力される各副搬送波は、加算器３により合成される。

ガード区間挿入器２１は、シンボル毎に、加算器３から出力される合成信号に対して、ガード区間（Guard Interval）を挿入する。ここで、このガード区間は、無線伝送路のマルチパスによる影響を排除するために挿入される。なお、図１に示した既存の送信装置のガード区間挿入器４は、予め固定的に決められたガード区間を挿入するが、実施形態のガード区間挿入器２１は、送信装置と受信装置との間の通信状態に応じて決められるガード区間を挿入する。なお、ガード区間の長さは、ガード区間制御部（ＧＩＮＳＣＮＴ：Guard Interval Control Unit）２３により、信号系列ごとに決定される。

利得調整器２２は、例えば乗算器であり、ガード区間が挿入された信号に利得係数αを乗算する。これにより、送信すべき信号の振幅または電力が調整される。なお、利得係数αは、基本的に、信号系列ごとに挿入されるガード区間の長さに対応して決定される。

上述のようにして得られる各信号系列ごとの合成信号は、図１に示した既存の送信装置と同様に、加算器（ＡＤＤ）５により加算される。そして、加算器５の出力は、送信機（ＴＸ）６により所定の高周波信号に変換された後、アンテナ７を介して送信される。

このように、実施形態の送信装置では、送信すべき信号系列（Ｓi、Ｓj）ごとに、送信装置と受信装置との間の通信状態に応じて決められるガード区間が挿入される。また、送信すべき信号系列（Ｓi、Ｓj）ごとに、挿入されたガード区

間の長さに対応して送信信号の振幅または電力が調整される。

図７は、本発明の実施形態の受信装置の構成図である。ここでは、この受信装置は、図６に示す送信装置により送信された無線信号から信号系列Ｓｉを受信するものとする。なお、この受信装置は、図５においては、例えば、移動機に相当する。また、図７では、信号を受信するために必要な周波数同期機能、およびタイミング同期機能などは省略されている。

アンテナ１１により受信された信号は、受信機（ＲＸ）１２によりベースバンド信号Ｓrx に変換された後、副搬送波復調器（ＦＤＥＭ：Frequency Demodulator）１３によりm個の受信信号列に変換される。ここで、副搬送波復調器１３は、m個の入力端子を備えており、それらの入力端子には、同一のベースバンド信号Ｓrx が並列に入力される。そして、副搬送波復調器１３は、ベースバンド信号Ｓrx に対してそれぞれ角周波数ω1～ωm を持った周期波を乗算することにより、各副搬送波ごとに信号を復調する。なお、副搬送波復調器１３の処理は、例えば、フーリエ変換演算により実現される。

ガード区間削除器３１は、ガード区間制御部（ＧＣＮＴｉ：Guard Interval Control Unit）３２からの指示に従って、各受信信号列からそれぞれガード区間を削除する。なお、ガード区間制御部３２は、送信装置において信号系列Ｓｉに対して挿入されたガード区間の長さを認識しており、その値をガード区間削除器３１に通知する。したがって、ガード区間削除器３１は、送信装置で挿入されたガード区間を適切に除去することができる。

拡散復調器１５は、各受信信号系列を逆拡散するために、送信装置において使用された拡散符号と同じ拡散符号Ｃｉを各受信信号列にそれぞれ乗算する。そして、加算器１６を用いて拡散復調器１５から出力される各信号を加算することにより、信号系列Ｓｉが再生される。

図８および図９は、実施形態のＯＦＤＭ伝送システムにおける伝送信号の例

である。ここで、図８は、最大伝送遅延差の小さい位置にいる移動機（受信装置）へ送信すべき伝送信号を模式的に示しており、図９は、最大伝送遅延差の大きい位置にいる移動機へ送信すべき伝送信号を模式的に示している。なお、図８に示す伝送信号のシンボル周期が「Ｔ1」であるのに対し、図９に示す伝送信号のシンボル周期が「T2」であるが、これらの周期は互いに同じであってもよいし、互いに異なっていてもよい。

最大伝送遅延差の小さい位置にいる移動機へ信号を送信する場合は、図８に示すように、各副搬送波ごとに、シンボル周期Ｔ1 に対してガード区間Ｔg1 が挿入される。したがって、信号は、区間Ｔs1 を利用して伝送される。一方、最大伝送遅延差の大きい位置にいる移動機へ信号を送信する場合は、図９に示すように、各副搬送波ごとに、シンボル周期Ｔ2 に対してガード区間Ｔg2 が挿入される。したがって、信号は、区間Ｔs2 を利用して伝送される。そして、このとき、ガード区間Ｔg1 は、ガード区間Ｔg2 よりも短く設定される。すなわち、ガード区間の長さは、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差が大きくなると、それに応じて長くなる。

また、最大伝送遅延差の小さい位置にいる移動機へ信号を送信する場合は、図８に示すように、信号の送信電力は「Ｐ1」に制御される。一方、最大伝送遅延差の大きい位置にいる移動機へ信号を送信する場合は、図９に示すように、信号の送信電力は「Ｐ2」に制御される。ここで、電力Ｐ2 は、電力Ｐ1 よりも大きい。すなわち、信号の送信電力は、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差が大きくなると、それに応じて大きくなる。

続いて、ガード区間の挿入／除去について説明する前に、ガード区間そのものについて簡単に説明をする。

図１０は、ガード区間について説明するための図であり、受信装置が受信した信号の波形が模式的に示されている。ここで、実線ａは、受信装置に最初に

到着した信号（基準波）の波形を表し、破線ｂは、受信装置に到着した遅延信号（遅延波）の波形を表している。なお、図１０では、１つの遅延波のみが描かれているが、実際には、通常、２以上の遅延波が存在する。

図１０において、時刻Ｔ１以前は、基準波および遅延波がそれぞれ連続したサイン波なので、受信装置は、それらの合成波から対応するシンボル情報を再生することができる。しかし、シンボル情報が「＋１」から「－１」に変化したとき、あるいは「－１」から「＋１」に変化したときは、そのシンボル情報を伝送する信号の位相が転移する。図１０に示す例では、時刻Ｔ１において基準波の位相が転移しており、時刻Ｔ2において遅延波の位相が転移している。すなわち、この場合、時刻Ｔ１と時刻Ｔ2との間の期間では、基準波は位相転移後の情報を伝送しており、遅延波は位相転移前の情報を伝送していることになる。したがって、この期間は、一方の信号波が他方の信号波に対する干渉波となり、受信波からシンボル情報を適切に再生することができないことがある。

上記干渉による影響は、例えば、図１０に示す例では、受信波から信号を再生する際に、時刻Ｔ1と時刻Ｔ2との間の受信波を使用しないことにより回避される。そして、ＯＦＤＭ通信システムでは、この期間を含む所定の期間をガード区間として定義し、受信装置において信号再生が行われないようにしている。したがって、ガード区間の長さは、最初に到着する信号波と最後に到着する遅延波との遅延差（最大伝送遅延差）よりも大きく設定される必要がある。

ところが、上述したように、最大伝送遅延差は、送信装置と受信装置との間の距離などにより変化する。したがって、実施形態の通信システムでは、ガード区間の長さが最大伝送遅延差に対応して決定されるようになっている。

次に、送信装置においてガード区間を挿入する方法を説明する。ここでは、図６に示す副搬送波変調器２の処理は、逆フーリエ変換演算により実現されるものとする。

図１１は、副搬送波変調器２により実行される逆フーリエ変換を説明する図である。ここでは、シンボル周期を「Ｔ」、シンボル周期ごとに挿入されるガード区間を「Tg」、シンボル周期ごとの信号時間を「Ts（＝Ｔ－Tg)」とする。

副搬送波変調器２には、上述したように、拡散変調器１から出力されるｍ個の情報が入力される。ここで、各情報は、それぞれ対応する周波数の副搬送波に割り当てらる。すなわち、副搬送波変調器２は、周波数軸上に配置されたｍ個の信号を受ける。そして、この周波数軸上のｍ個の信号は、図１１に示すように、シンボル周期Ｔごとに実行される逆フーリエ変換により、時間軸上のｍ個のサンプルから構成される信号系列に変換される。このとき、時間軸上のｍ個のサンプルは、信号時間Ts内に配置される。

図１２は、ガード区間を挿入する処理を説明する図である。ガード区間挿入器２１は、信号時間Ts内に配置されたｍ個のサンプルを受け取ると、ガード区間Tgに相当する個数のサンプル成分を信号時間Tsの末尾から抽出し、それらを信号時間Tsの直前に複写する。図１２に示す例では、ガード区間Tgが３サンプリング時間に相当し、ｍ個のサンプル「１」～「ｍ」のうちから、「ｍ－２」「ｍ－１」「ｍ」が抽出されて信号時間Ts の直前に複写されている。そして、この複写により、シンボル時間Ｔ（＝Tg＋Ts）の時間軸上の信号系列が作成される。

図１３は、ガード区間を挿入する処理を実現する構成の実施例である。上述したように、副搬送波変調器２は、逆フーリエ変換器によって実現され、シンボル周期ごとに、周波数軸上のｍ個の信号を時間軸上のｍ個のサンプル（ｔ1～ｔm）に変換する。そして、ガード区間挿入器２１は、まず、ガード区間Tgにおいて、「ｔm-2」「ｔm-1」「ｔm」を順番に読み出して出力し、それに続く信号時間Ts において「ｔ1」～「ｔm」を順番に読み出して出力していく。これにより、ガード区間が挿入された信号系列が作成される。

上記構成においてガード区間の長さは、「信号時間Ts の前に出力するサンプルの数」を変えることにより制御される。この場合、ガード区間Tg、信号時間Ts、および逆フーリエ変換の周期（すなわち、シンボル周期T）が所定の関係（T＝Tg＋Ts）を満たすように、サンプル値の読出し間隔が決定される。

一例を示す。ここでは、シンボル周期＝T、ガード区間Tg＝０．２T、信号時間Ts＝０．８T、副搬送波の多重数m＝１０００であるものとする。この場合、ガード区間挿入器２１には、シンボル周期ごとに、時間軸上の１０００個のサンプル（t 1～ t 1000）が入力される。そして、まず、２５０（＝１０００×０．２÷０．８）個のサンプル（t 751～ t 1000）を読み出して出力する。続いて、上記１０００個のサンプル（t 1～ t 1000）を読み出して出力する。このとき、サンプル値の読出し間隔は、「T／１２５０」である。また、ガード区間Tg＝０．１T、信号時間Ts＝０．９T、副搬送波の多重数m＝１０００であるものとすると、ガード区間挿入器２１は、まず、１１１（＝１０００×０．１÷０．９）個のサンプル（t 890～ t 1000）を読み出して出力し、それ続いて、上記１０００個のサンプル（t 1～ t 1000）を読み出して出力する。このとき、サンプル値の読出し間隔は、「T／１１１１」である。

なお、実施形態では、複数の副搬送波が合成された後にガード区間が挿入されているが、原理的には、副搬送波ごとにガード区間を挿入することも可能である。

図１４は、受信装置において受信波からガード期間を削除する処理を実現する構成の実施例である。ここでは、図１１～図１３に示すようにして作成された信号列（t m-2, t m-1, t m, t 1, t 2, t 3,... t m）が受信されるものとする。なお、図７に示す受信装置では、副搬送波変調を行った後にガード区間が削除されるように描かれているが、図１４に示す構成では、これらの処理は一体的に実行される。

ガード区間削除器３１は、スイッチ４１およびシフトレジスタ４２を備える。そして、信号系列（ｔm-2，ｔm-1，ｔm，ｔ1，ｔ2，ｔ3，...ｔm）を受信すると、スイッチ４１を適切にＯＮ／ＯＦＦ制御することにより、ガード区間に配置されている所定数のサンプル値（ここでは、ｔm-2，ｔm-1，ｔm）を廃棄し、後続のm個のサンプル値（ｔ1～ｔm）をシフトレジスタ４２に送る。ここで、ガード区間削除器３１は、送信装置において挿入されたガード区間の長さ（あるいは、ガード区間内のサンプル数）を認識しており、それに基づいてスイッチ４１のＯＮ／ＯＦＦ状態を制御する。一方、副搬送波復調器１３として動作するフーリエ変換器は、シフトレジスタ４２にm個のサンプル値が蓄積されると、それらのサンプル値についてフーリエ変換を行うことにより、副搬送波ごとの信号ｆ1～ｆm を得る。なお、この処理は、シンボル周期Ｔごとに繰り返し実行される。

このように、実施形態のセルラ通信システムでは、送信装置（基地局）から受信装置（移動機）へ信号を送信する際、それらの間の最大伝送遅延差に基づいて、ガード区間の長さ、および送信電力が決定される。ここで、送信装置と受信装置との間の距離が短い場合は、最大伝送遅延差が小さくなり、ガード区間が短くなる。そして、ガード区間が短くなると、それに応じて受信装置において信号再生に寄与する信号時間が長くなるので、送信電力を低くすることができる。したがって、システム全体として干渉電力が減少し、伝送容量が増加することになる。

次に、上述の送信装置および受信装置の実施例を説明する。

第１の実施例：

図１５および図１６は、第１の実施例の送信装置および受信装置の構成図である。これらの装置の基本構成は、それぞれ、図６に示した送信装置および図７に示した受信装置と同じである。ただし、第１の実施形態の送信装置は、時

間多重された複数の信号系列を１つのＯＦＤＭ－ＣＤＭユニット（拡散変調器１、副搬送波変調器２、加算器３、ガード区間挿入器２１）により一括して変調することができる。

すなわち、信号系列Ｓi1 および信号系列Ｓi2 は、図１７に示すように、時間多重化部（TDMi）５１により多重化される。ここでは、これらの信号系列は、互いに異なる最大伝送遅延差を有する回線を介して伝送されるものとする。そして、この信号系列は、拡散変調器１および副搬送波変調器２により変調された後、ガード区間挿入器２１に与えられる。

ガード区間挿入器２１は、入力される信号系列に対して、対応する最大伝送遅延差よりも広いガード区間を挿入する。ここで、各信号系列に対するガード区間は、ガード区間制御部２３により設定される。また、利得調整器２２は、挿入されたガード区間に応じて決まる利得係数αを送信信号に乗算する。具体的には、図１７に示す例では、信号系列Ｓi1 が入力されている期間は、シンボル周期ごとにガード区間Ｔg1 が挿入され、信号の送信電力が「Ｐ1」になるように利得係数$\alpha i(t)$が制御される。一方、信号系列Ｓi2 が入力されている期間は、シンボル周期ごとにガード区間Ｔg2 が挿入され、信号の送信電力が「Ｐ2」になるように利得係数$\alpha i(t)$が制御される。

そして、上述のようにして変調された信号は、他の系の信号と合成された後、アンテナ７を介して送信される。

受信装置の基本的な動作は、図７を参照しながら説明した通りである。ただし、この受信装置は、自分宛ての信号のみを再生する。例えば、信号系列Ｓi1 および信号系列Ｓi2 が時間多重された信号から信号系列Ｓi1 を再生する場合には、ガード区間制御部３２は、信号系列Ｓi1 を受信している期間に、ガード区間Tg1 を削除するようにガード期間削除器３１に対して指示を与える。そして、ガード区間削除器３１は、その指示に従って信号系列Ｓi1 のシンボル周期ごと

にガード区間を削除する。このとき、信号系列Si2 を受信している期間は、ガード区間は削除される必要はない。

ガード区間削除器３１の出力は、拡散復調器１５により逆拡散復調される。このとき、拡散復調器１５は、ガード区間Tg1 が削除された信号時間Ts1 について逆拡散復調を行う。そして、分離部（DML）５２は、復調された信号から、信号系列Si1 に対応する時間スロットにおいてデータを出力する。

このように、第１の実施例の通信システムでは、時間多重された複数の信号系列を１つのOFDM－CDMユニット（拡散変調器１、副搬送波変調器２、加算器３、ガード区間挿入器２１）により一括して変調できる。

第２の実施例：

第２の実施例の通信システムは、第１の実施例の通信システムの変形例である。すなわち、第１の実施例のシステムでは、時間多重された信号系列Si1 および信号系列Si2 がOFDM－CDMを利用して伝送される。ここで、信号系列Si1 および信号系列Si2 は、基本的に、それぞれ対応する移動機に送信されることを想定している。これに対して、第２の実施例のシステムでは、時間多重された報知情報Bi および信号系列Si1 がOFDM－CDMを利用して伝送される。ここで、信号系列Si1 は、所定の１または複数の受信装置に対して送信されるが、報知情報Biは、サービスエリア内のすべての受信装置（移動機）に対して送信される。したがって、この報知情報Biは、サービスエリア内の最も遠くに位置する受信装置（すなわち、最大伝送遅延差が最も大きくなる受信装置）に適切に伝送されるようなガード区間が設定され、且つ、送信電力が決定される必要がある。

図１８および図１９は、第２の実施例の送信装置および受信装置の構成図である。これらの装置の基本構成は、それぞれ、図１５に示した送信装置および図１６に示した受信装置と同じである。

第２の実施例では、ガード区間挿入器２１は、図２０に示すように、ガード区間制御部２３からの指示に従って、報知情報Biが入力されている期間は、シンボル周期ごとにガード区間Tg1を挿入し、信号系列Si1が入力されている期間は、シンボル周期ごとにガード区間Tg2 を挿入する。ここで、報知情報Biに対して挿入されるガード区間Tg1 は、サービスエリア内において生じる最も大きな最大伝送遅延差よりも長くなるように設定される。例えば、図５において、基地局から移動機ＭＳ１～ＭＳ３へ報知情報を送信する際、基地局から移動機ＭＳ３への回線の最大伝送遅延差が最も大きかったとすると、ガード区間Tg1 の長さは、その最大伝送遅延差よりも長くなるように設定される。一方、信号系列Si1 に対して挿入されるガード区間Tg2 は、対応する受信装置への回線の最大伝送遅延差よりも長くなるように設定される。例えば、図５において、基地局から移動機ＭＳ１信号系列Si1 を送信する際には、ガード区間Tg2 の長さは、基地局から移動機ＭＳ１への回線の最大伝送遅延差よりも長くなるように設定される。

また、利得調整器２２は、ガード区間挿入器２１により挿入されたガード区間に応じた利得係数αを送信信号に乗算する。具体的には、図２０に示す例では、利得係数αi(t)は、報知情報Biを伝送するための信号の送信電力が「P1」となり、信号系列Si1を伝送するための信号の送信電力が「P2」になるように制御される。したがって、このように制御される利得係数αを送信信号に乗算することにより、報知情報Biはサービスエリア内のすべての受信装置に伝送されるように大きな送信電力で送信され、信号系列Si1 は対応する受信装置に伝送される範囲で必要最小限の送信電力で送信される。

受信装置では、ガード区間制御部３２は、報知情報Biを受信している期間はガード区間Tg1を指示し、信号系列Si1を受信している期間はガード区間Tg2を指示する。そして、ガード区間削除器３１は、ガード区間制御部３２からの

指示に従って受信信号からガード区間を削除する。さらに、ガード区間が削除された信号は、拡散復調器１５により逆拡散された後、分離部５２により報知情報Biおよび信号系列Si1に分離される。

なお、報知情報Biに対して挿入されるガード区間Tg1の長さは、例えば、以下のようにして決定される。

（１）通信エリアの大きさに基づいて決定する。すなわち、送信装置がカバーする通信エリアの大きさに基づいて、報知情報Biが最も遅延して到着する受信装置までの遅延時間を推定し、その遅延時間に従ってガード区間Tg1の長さを決定する。

（２）報知情報Bi を送信する際の送信装置の送信電力に基づいて決定する。すなわち、報知情報Bi の送信電力により、その報知情報Bi を複数の受信装置に送信する際の伝送遅延時間の最大値を推定し、その遅延時間に従ってガード区間Tg1の長さを決定する。

（３）通信エリア内に存在する複数の受信装置との間の通信環境に基づいて決定する。すなわち、送信装置がカバーする通信エリア内に存在する複数の受信装置との間の通信環境をそれぞれ求め、これに基づいてガード区間Tg1の長さを決定する。具体的には、通信環境が最も厳しい受信装置に合わせてガード区間Tg1の長さを決定する。

（４）通信エリア内の最大遅延時間に基づいて決定する。すなわち、送信装置から通信エリア内に存在する複数の受信装置へ報知情報Bi を送信したときの遅延時間を受信装置ごとに測定し、それらのうちの最大遅延時間に基づいてガード区間Tg1の長さを決定する。

第３の実施例：

第３の実施例の通信システムでは、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差を検出し、その検出結果に基づいてガード区間および

送信電力が決定される。したがって、第３の実施例における送信装置および受信装置は、そのための機能を備えている。

図２１は、第３の実施例の送信装置の構成図である。この送信装置は、対応する受信装置において検出された最大伝送遅延差を表す最大伝送遅延差情報（τ）を受け取り、その情報に基づいてガード区間および送信電力を決定する機能を備えている。即ち、ガード区間制御部（ＧＩＮＳＣＮＴ）６１は、対応する受信装置において検出された最大伝送遅延差に基づいて、挿入すべきガード区間の長さを決定する。具体的には、ガード区間制御部６１ｉは、信号系列Ｓi1 および／または信号系列Ｓi2 を受信する受信装置から送られてくる最大伝送遅延差情報（τi）に基づいて、信号系列Ｓi1 および／または信号系列Ｓi2 を伝送するための信号に挿入すべきガード区間を決定する。また、電力制御部（ＰＣＮＴ）６２は、対応する受信装置において検出された最大伝送遅延差に基づいて、利得係数αを決定する。具体的には、電力制御部６２ｉは、信号系列Ｓi1 および／または信号系列Ｓi2 を受信する受信装置から送られてくる最大伝送遅延差情報（τi）に基づいて、信号系列Ｓi1 および／または信号系列Ｓi2 を伝送するための信号に乗算すべき利得係数αを決定する。

そして、ガード区間挿入器２１は、シンボル周期ごとに、送信信号に対してガード区間制御部６１により決定されたガード区間を挿入する。また、利得調整器２２は、電力制御部６２により決定された利得係数αを送信信号に乗算することにより、ガード区間の長さに対応する送信電力を実現する。

図２２は、第３の実施例の受信装置の構成図である。この受信装置は、送信装置から送られてきた信号の最大伝送遅延差を検出する機能を備えている。すなわち、遅延差検出部（ＤＭＥＳ）６３は、受信したベースバンド信号Ｓrx から最大伝送遅延差を検出し、その検出結果を表す最大伝送遅延情報をガード区間制御部６４および対応する送信装置に通知する。ガード区間制御部６４は、

遅延差検出部６３からの通知に従ってガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図２３は、図２２に示す遅延検出部６３の一例の構成図である。遅延差検出部６３は、ベースバンド信号Ｓrx を時間Ｔs だけ遅延させる遅延回路７１、乗算器７２ａおよび積分器７２ｂから構成される相関検出回路７２、相関検出回路７２により検出された相関値と予め決められている所定のしきい値とを比較する比較回路７３、および比較回路７３による比較結果に基づいて最大伝送遅延差を検出する検出回路７４を含む。ここで、乗算器７２ａは、ベースバンド信号Ｓrx にその遅延信号を乗算し、積分器７２ｂは、乗算器７２ａの出力を積分する。以下、図２４を参照しながら遅延差検出部６３の動作を説明する。

相関検出回路７２には、ベースバンド信号Ｓrx およびそのベースバンド信号Ｓrx を時間Ｔs だけ遅延させた信号（遅延信号）が入力される。ここで、各シンボル周期内のガード区間Ｔg には、図１１〜図１３を参照しながら説明したように、信号時間Ｔs の最後尾部分のサンプル値が複写されている。このため、ベースバンド信号Ｓrx とその遅延信号との間では、ベースバンド信号Ｓrx の最後尾部分と遅延信号のガード区間とが重なったときに相関（自己相関）が高くなる。ただし、送信装置と受信装置との間に伝送遅延の異なる複数のパスが存在する場合には、各パスを介して信号を受信するごとに相関値のピークが発生する。したがって、比較回路７３を用いて上記相関値と予め設定されているしきい値とを比較すれば、各パスを介して信号を受信したタイミングをそれぞれ検出できる。よって、最初に信号を受信したタイミングと、最後に信号を受信したタイミングとの時間差を測定することにより、最大伝送遅延差が検出される。例えば、図４に示す通信環境においては、図２５に示すようにして最大伝送遅延差が検出される。

このように、第３の実施例では、送信装置と受信装置との間の回線の最大伝送遅延差が測定され、その結果に基づいてガード区間が挿入／削除されるので、ガード区間の幅を動的に変化させることが可能である。また、上記最大伝送遅延差の測定結果に従って送信信号の利得係数が決定されるので、常に、送信電力を必要最小限に抑えられる。

第４の実施例：

第４の実施例の通信システムでは、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送信電力が決定される。したがって、第４の実施例における送信装置および受信装置は、そのための機能を備えている。

図２６は、第４の実施例の送信装置の構成図である。この送信装置は、対応する受信装置との間の伝送距離の推定値を表す伝送距離情報（L）を受け取り、その情報に基づいてガード区間および送信電力を決定する機能を備えている。即ち、ガード区間制御部（ＧＩＮＳＣＮＴ）８１は、送信装置と受信装置との間の伝送距離に基づいて、挿入すべきガード区間の長さを決定する。具体的には、ガード区間制御部８１ｉは、信号系列Ｓi1 および／または信号系列Ｓi2 を受信する受信装置から送られてくる伝送距離情報（Li）に基づいて、信号系列Ｓi1 および／または信号系列Ｓi2 を伝送するための信号に挿入すべきガード区間を決定する。また、電力制御部（ＰＣＮＴ）８２は、上記伝送距離に基づいて、利得係数αを決定する。具体的には、電力調整部８２ｉは、信号系列Ｓi1 および／または信号系列Ｓi2 を受信する受信装置から送られてくる伝送距離情報（Li）に基づいて、信号系列Ｓi1 および／または信号系列Ｓi2 を伝送するための信号に乗算すべき利得係数αを決定する。

そして、ガード区間挿入器２１は、シンボル周期ごとに、送信信号に対してガード区間制御部８１により決定されたガード区間を挿入する。また、利得調

整器２２は、電力制御部８２により決定された利得係数αを送信信号に乗算することにより、ガード区間の長さに対応する送信電力を実現する。

図２７は、第４の実施例の受信装置の構成図である。この受信装置は、送信装置と当該受信装置との間の伝送距離を推定する機能を備えている。すなわち、距離推定部（ＬＭＥＳ）８３は、受信したベースバンド信号Ｓrx に基づいて送信装置と当該受信装置との間の伝送距離を推定し、その推定結果を表す伝送距離情報Ｌをガード区間制御部８４および対応する送信装置に通知する。ガード区間制御部８４は、距離推定部８３からの通知に従ってガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図２８は、図２７に示す距離推定部８３の一例の構成図である。距離推定部８３は、第３の実施例において説明した遅延差検出部６３および変換テーブル８５から構成される。

送信装置と受信装置との間の伝送距離は、その間の回線の最大伝送遅延差と相関があり、伝送距離が長くなるほど最大伝送遅延差も大きくなることが知られている。したがって、これらの間の関係を実験またはシミュレーション等により予め求めておけば、最大伝送遅延差を検出することによって伝送距離を推定することができる。このため、距離推定部８３の変換テーブル８５には、伝送距離と最大伝送遅延差との関係を表す情報が格納されている。そして、遅延差検出部６３により検出された最大伝送遅延差をキーとしてその変換テーブル８５を検索することにより、送信装置と受信装置との間の伝送距離が推定される。

第５の実施例：

第５の実施例の通信システムでは、第４の実施例と同様に、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送

信電力が決定される。ただし、第５の実施例における推定方法は、第４の実施例のそれと異なっている。

図２９は、第５の実施例の送信装置の構成図である。この送信装置は、対応する受信装置からタイミング情報（Ｔ）を受け取ってそれに基づいて送信装置と受信装置との間の伝送距離を推定する機能、およびその伝送距離の推定値に基づいてガード区間および送信電力を決定する機能を備えている。

ガード区間制御部（ＧＩＮＳＣＮＴ）９１または電力制御部（ＰＣＮＴ）９２は、対応する受信装置から送られてくるタイミング信号Ｔに基づいて、当該送信装置と対応する受信装置との間の距離を推定する。すなわち、第５の実施例では、送信装置から信号が送信され、その信号が対応する受信装置により検出され、さらにその受信装置において上記信号が検出された旨が送信装置に通知される。ここで、上記信号が上記受信装置において検出されたタイミングは、タイミング情報Ｔを用いて送信装置に通知される。したがって、ガード区間制御部９１または電力制御部９２は、信号を送信したときから、対応する受信装置からタイミング情報Ｔを受信するまでの時間をモニタすることにより、送信装置と受信装置との間の伝送時間および伝送距離を推定できる。そして、上記伝送距離の推定値は、伝送距離情報Ｌを利用して対応する受信装置に送られる。

なお、ガード区間制御部９１は、伝送距離の推定値に基づいてガード区間の長さを決定する。また、電力制御部９２は、伝送距離の推定値に基づいて利得係数αを決定する。これらの処理は、基本的に、第４の実施例と同じである。

図３０は、第５の実施例の受信装置の構成図である。この受信装置は、送信装置から送出された信号の受信タイミングを検出する機能を備えている。すなわち、タイミング生成部（ＴＧＥＮ）９３は、受信したベースバンド信号S_{rx}を基準として受信タイミングを検出し、タイミング信号Ｔを生成する。そして、生成したタイミング信号Ｔは、送信装置へ送られる。また、ガード区間制御部

（ＧＣＮＴ）９４は、送信装置から送られてくる伝送距離情報Ｌに基づいてガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３１は、図３０に示すタイミング生成部９３の一例の構成図である。タイミング生成部９３は、第３の実施例において説明した遅延回路７１、相関検出回路７２、および最大値判定回路９５を含む。

上述したように、受信信号とその遅延信号との自己相関をモニタした場合、ガード区間を受信している期間の相関値が高くなる。したがって、その相関値をモニタすることにより、ガード区間の位置を検出できる。具体的には、最大値判定部９５を用いてシンボル周期ごとに上記相関値の最大値を検出することにより、ガード区間のタイミング（または、ガード区間の直後に相当するタイミング）を検出できる。そして、タイミング生成部９３は、検出したタイミングを表すタイミング情報Ｔを生成し、それを送信装置へ送る。

第６の実施例：

第６の実施例の通信システムでは、第４または第５の実施例と同様に、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送信電力が決定される。ただし、第６の実施例における推定方法は、第４または第５の実施例のそれと異なっている。

第６の実施例の通信システムでは、信号系列Ｓi1 および信号系列Ｓi2 を送信する際に、それらの系列にそれぞれ既知情報ＳＷが時間多重される。一方、受信装置は、受信信号に中に含まれている既知情報ＳＷを検出すると、その検出タイミングを送信装置に通知する。そして、送信装置は、既知情報ＳＷを送信したタイミングおよび対応する受信装置から送られてくるタイミング情報に基づいて、当該送信装置と受信装置との間の信号の伝送時間を検出し、その伝送時間から伝送距離を推定する。

図３２は、第６の実施例の送信装置の構成図である。この送信装置は、送信信号系列に既知情報ＳＷを多重化する機能、対応する受信装置からタイミング情報（Ｔ）を受け取ってそれに基づいて送信装置と受信装置との間の伝送距離を推定する機能、およびその伝送距離の推定値に基づいてガード区間および送信電力を決定する機能を備えている。

時間多重化部（TDM）５１は、信号系列Si1、Si2 を送信する際に、それらの系列にそれぞれ既知情報ＳＷを多重する。ここで、既知情報ＳＷは、特に限定されるものではないが、対応する受信装置がそのデータパターンを認識している必要がある。

ガード区間制御部（ＧＩＮＳＣＮＴ）１０１または電力制御部（ＰＣＮＴ）１０２は、対応する受信装置から送られてくるタイミング信号Ｔに基づいて、当該送信装置と対応する受信装置との間の距離を推定する。そして、この伝送距離の推定値は、伝送距離情報Ｌを利用して対応する受信装置に送られる。なお、伝送距離を推定する方法については後述する。

なお、ガード区間制御部１０１は、伝送距離の推定値に基づいてガード区間の長さを決定する。また、電力制御部１０２は、伝送距離の推定値に基づいて利得係数αを決定する。これらの処理は、基本的に、第４または第５の実施例と同じである。

図３３は、第６の実施例の受信装置の構成図である。この受信装置は、受信波から既知情報ＳＷを分離して出力する機能、および既知情報を受信した旨を送信装置に通知する機能を備えている。すなわち、タイミング生成部（ＴＧＥＮ）１０３は、分離部（ＤＭＬ）５２から出力された既知情報ＳＷを検出すると、その検出タイミングから所定時間経過後にタイミング信号Ｔを生成して送信装置へ送出する。また、ガード区間制御部（ＧＣＮＴ）１０４は、送信装置から送られてくる伝送距離情報Ｌに基づいてガード区間を決定し、それをガー

ド区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３４は、図３３に示すタイミング生成部１０３の一例の構成図である。タイミング生成部１０３には、当該受信装置により復調された信号列が入力される。ここで、この信号列は、送信装置において挿入された既知情報ＳＷを含んでいる。そして、この信号列は、既知情報ＳＷのワード長と等しい長さのシフトレジスタ１０５に順番に入力されていく。論理反転回路１０６、加算回路１０７、および比較回路１０８は、シフトレジスタ１０５に新たなデータの書き込まれるごとに、保持されているデータが既知情報ＳＷと一致するか否かを調べる。なお、論理反転回路１０６は、既知情報ＳＷのワードパターンに対応して設けられている。また、加算回路１０７は、シフトレジスタ１０５に保持されている各エレメントの値またはシフトレジスタ１０５に保持されている各エレメントの値の論理反転値を加算する。そして、比較回路１０８は、加算回路１０７による加算結果と予め設定されている閾値とを比較し、加算結果の方が大きかったときにタイミング信号Ｔを出力する。

このように、第６の実施例の通信システムでは、送信装置から受信装置へ既知情報ＳＷが送信され、その既知情報ＳＷを検出した旨が受信装置から送信装置へ通知される。したがって、送信装置から受信装置へ信号が伝送される際の伝送時間を「Ｔ1」、受信装置が既知情報ＳＷを検出してからタイミング情報を送信するまでの時間を「Ｔd」、受信装置から送信装置へタイミング情報が伝送される際の伝送時間を「Ｔ2」、既知情報ＳＷを送信してからタイミング情報を受信するまでの時間を「Ｔ0」とすると、下記の式が成立する。なお、「Ｔ2」は「Ｔ1」に比例するものとし、その比例定数を「β」とする。

$$
\begin{aligned}
T1 &= T0 - Td - T2 \\
&= T0 - Td - \beta \cdot T1
\end{aligned}
$$

$\therefore\ T1 = (T0 - Td) / (1 + \beta)$

ここで、送信装置と受信装置との間の伝送距離は、送信装置から受信装置へ信号が伝送される際の伝送時間（T1）に比例する。また、受信装置が既知情報ＳＷを検出してからタイミング情報を送信するまでの時間（Td）は既知である。したがって、送信装置は、既知情報ＳＷを送信してからタイミング情報を受信するまでの時間（T0）を測定することにより、送信装置と受信装置との間の伝送距離を推定できる。なお、この実施例では、ガード区間制御部１０１または電力制御部１０２がその伝送距離を推定する。

第７の実施例：

第７の実施例の通信システムでは、ガード区間の長さを変えながら伝送エラー率が測定され、所定の伝送品質が確保されるようにガード区間の長さ（および、送信電力）が決定される。したがって、第７の実施例における送信装置および受信装置は、そのための機能を備えている。

図３５は、第７の実施例の送信装置の構成図である。この送信装置は、既知パターンデータ（ＰＬj）を変調して送信する機能、および対応する受信装置から最大伝送遅延差情報（τ）を受け取ってそれに基づいてガード区間および送信電力を決定する機能を備えている。

既知パターンデータ（ＰＬj）は、拡散変調器１により拡散された後、副搬送波変調器２により変調される。ここで、既知パターンデータ（ＰＬj）は、特に限定されるものではないが、各受信装置により認識されているものとする。また、拡散変調器１は、既知パターンデータ（ＰＬj）に対応する拡散符号Ｃ（ＰＬj）により拡散される。

ガード区間挿入器（ＧＩＮＳj）２１は、シンボル周期ごとに、既知パターンデータ（ＰＬj）を伝送するための信号系列に比較的長いガード区間を挿入する。ここで、このガード区間は、例えば、サービスエリア内の最も遠い位置にいる

移動機（受信装置）へ信号を送信する場合を想定して決定されるようにしてもよい。また、利得調整器（Gj）２２は、ガード区間が挿入された信号系列が十分に大きな送信電力で送信されるように適切な利得係数αjを乗算する。ここで、この利得係数αjは、例えば、サービスエリア内の最も遠い位置にいる移動機（受信装置）へ信号を送信する場合を想定して決定されるようにしてもよい。そして、既知パターンデータ（ＰＬj）は、信号系列Si1、Si2 と合成されて送信される。

ガード区間制御部（ＧＩＮＳＣＮＴ）６１および電力制御部（ＰＣＮＴ）６２の動作は、第３の実施例において説明した通りである。すなわち、ガード区間制御部６１は、対応する受信装置から送られてくる最大伝送遅延差情報に基づいて、挿入すべきガード区間の長さを決定する。また、電力制御部６２は、対応する受信装置から送られてくる最大伝送遅延差情報に基づいて、利得係数αを決定する。

図３６は、第７の実施例の受信装置の構成図である。この受信装置は、既知パターンデータ（ＰＬj）を抽出してその伝送エラーを測定する機能、および伝送エラー率に基づいて最大伝送遅延差情報を生成する機能を備えている。

受信波は、復調回路により復調される。このとき、拡散復調器（ＳＤＥＭ）１５において、信号系列Si1 を復調するときは拡散符号Ciが使用され、既知パターンデータ（ＰＬj）を復調するときには拡散符号Ｃ（ＰＬj）が使用される。そして、分離部５２は、再生された信号列を、信号系列Si1 および既知パターンデータ（ＰＬj）に分離する。

遅延差検出部（ＤＭＥＳ）１１１は、再生された既知パターンデータ（ＰＬj）の伝送エラー率を測定し、その伝送エラー率に基づいて最大伝送遅延差情報τを生成する。この最大伝送遅延差情報τは、ガード区間制御部（ＧＣＮＴ）１１２に与えられると共に、送信装置に送られる。そして、ガード区間制御部１

１２は、その最大伝送遅延差情報τに基づいてガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３７は、図３６に示す遅延差検出部１１１の動作を示すフローチャートである。ここでは、予め複数のガード区間長データτ0～τn が用意されているものとする。また、ガード区間長データτ0～τn の中で、「τ0」が最小であり、「τn」が最大であるものとする。なお、このフローチャートの処理は、たとえば、既知パターンデータ（ＰＬj）を受信するごとに実行される。

ステップＳ１では、拡散復調器１５に拡散符号Ｃ（ＰＬj）を設定する。ここで、この拡散符号Ｃ（ＰＬj）は、送信装置において既知パターンデータ（ＰＬj）を拡散する際に使用されてものである。これにより、以降、受信信号が逆拡散されると、既知パターンデータ（ＰＬj）が再生されることになる。ステップＳ２では、ガード区間長データを指定するための変数を初期化する。すなわち、「ｉ＝０」が設定される。

ステップＳ３では、ガード区間制御部１１２にガード区間長データτｉを設定する。ただし、この時点では、「ｉ＝０」であるので、ガード区間制御部１２３には「ガード区間長データτ0」が設定されることになる。ここで、「ガード区間長データτ0」は、予め用意されている候補データの中で最も短い値を持っている。また、このとき、分離部５２は、再生された既知パターンデータ（ＰＬj）が遅延差検出部１１１に導かれるように出力する。

ステップＳ４では、再生された既知パターンデータ（ＰＬj）の誤り率（誤りビット数）を調べる。そして、この誤り率が予め設定されているしきい値よりも高かった場合には、十分な通信品質が得られていないものとみなし、ステップＳ５へ進む。ステップＳ５では、変数ｉをインクリメントできるか否かが調べられる。そして、可能であれば、ステップＳ６において変数ｉがインクリメ

ントされた後、ステップＳ３に戻る。

このように、ステップＳ３～Ｓ６では、ガード区間制御部１１２に設定すべきガード区間長を少しずつ長くしていきながら、それぞれについて既知パターンデータ（ＰＬj）の誤り率が測定される。そして、既知パターンデータ（ＰＬj）の誤り率がしきい値以下になった時点で、ステップＳ７へ進む。したがって、上記処理により、所望の通信品質が得られる範囲内で、できるかぎり短いガード区間長が決定される。なお、この時点で、ガード区間制御部１１２には、最適なガード区間が設定されていることになる。

ステップＳ７では、拡散復調器１５に拡散符号Ｃiを設定する。ここで、拡散符号Ｃiは、送信装置において信号系列Ｓi1、Ｓi2を拡散する際に使用されたものである。したがって、以降、拡散復調器１５は、受信信号から信号系列Ｓi1を復調できるようになる。ステップＳ８では、ステップＳ３～Ｓ６において決定されたガード区間長を送信装置に通知する。

このように、第７の実施例では、伝送エラー率を測定しながら所定の通信品質が確保されるようにガード区間の長さ（および、送信電力）が決定される。したがって、必要最小限のガード区間および送信電力で所望の通信品質が確保される。

第８の実施例：

第８の実施例の通信システムは、第７の実施例の通信システムの変形例である。すなわち、第７の実施例では、受信装置に設定すべきガード区間長が決定され、その値が送信装置に通知される構成であった。これに対して、第８の実施例では、受信装置に設定すべきガード区間長に基づいて送信装置と受信装置との間の伝送距離が推定され、その推定結果が送信装置に通知される。

図３８は、第８の実施例の送信装置の構成図である。この送信装置は、基本的には、図３５に示した第７の実施例の送信装置と同じである。ただし、第８

の実施例の送信装置は、図３５に示したガード区間制御部（ＧＩＮＳＣＮＴ）６１および電力制御部（ＰＣＮＴ）６２の代わりに、ガード区間制御部（ＧＩＮＳＣＮＴ）８１および電力制御部（ＰＣＮＴ）８２が設けられている。なお、ガード区間制御部８１および電力制御部８２の動作は、第４の実施例において説明した通りである。すなわち、ガード区間制御部８１は、対応する受信装置から送られてくる伝送距離情報Ｌに基づいて、挿入すべきガード区間の長さを決定する。また、電力制御部８２は、対応する受信装置から送られてくる伝送宇距離情報Ｌに基づいて、利得係数αを決定する。

図３９は、第８の実施例の受信装置の構成図である。この受信装置は、図３６に示した第７の実施例の受信装置の遅延差検出部１１１、ガード区間制御部１１２の代わりに、距離推定部（ＬＭＥＳ）１２１、変換テーブル（ＴＢＬ）１２２、ガード区間制御部（ＧＣＮＴ）１２３を備える。ここで、距離推定部１２１およびガード区間制御部１２３は、まず、第７の実施例と同様に、最適なガード区間長を決定する。その後、距離推定部１２１は、変換テーブル１２２にアクセスし、決定したガード区間長に対応する伝送距離を取得する。そして、その伝送距離を表す伝送距離情報Ｌを送信装置に通知する。なお、変換テーブル１２２は、図２８に示した変換テーブル８５に相当し、ガード区間長と伝送距離との対応関係が格納されている。

図４０は、図３９に示す距離推定部１２１の動作を示すフローチャートである。図４０において、ステップＳ１～Ｓ７は、図３７に示した第７の実施例における処理と同じである。すなわち、ステップＳ１～Ｓ７において、受信装置に設定すべきガード区間長τiが決定される。続いて、ステップＳ１１では、変換テーブル１１２を参照して、ガード区間長τiを伝送情報Liに変換する。そして、ステップＳ１２において、ステップＳ１１で取得した伝送情報を送信装置に通知する。

このように、本発明によれば、セルラ通信システムにおける基地局とそのサービスエリア内の移動機との間の伝送路で生じる最大伝送遅延差に応じてガード区間および送信電力が適切に設定されるので、干渉の発生が低減される。あるいは、伝送路の送信帯域内での伝送容量が最適化されるので、通信システムの効率的な運用が可能となり、総伝送容量を増加させることができる。

なお、ガード区間および送信電力は、送信装置と受信装置との間の回線の最大伝送遅延差（または、伝送距離）に応じて動的に制御されてもよいし、固定的に設定されてもよい。例えば、通信の開始時にガード区間および送信電力が決定され、以降、その通信が終了するまでそれらが変化しないようにしてもよい。また、通信中に、随時、ガード区間および送信電力が動的に調整されてもよい。さらに、送信装置および受信装置の位置が変化しない場合には、初期設定処理においてガード区間および送信電力が決定されてもよい。

また、本発明では、最大伝送遅延差（または、伝送距離）に応じてガード区間および送信電力が決定されるが、ガード区間長と送信電力の関係は、たとえば、実験またはシミュレーション等により予め一意に決められていてもよい。

請求の範囲

１．　直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

信号系列を用いて複数の副搬送波を変調する変調手段と、

上記変調手段の出力にガード区間を挿入する挿入手段と、

上記ガード区間が挿入された変調信号を送信する送信手段を有し、

上記受信手段は、

上記送信装置から送信された変調信号について副搬送波ごとにガード区間の削除処理と復調処理を行い、信号系列を再生する復調手段を有し、

上記ガード区間の長さは、上記送信装置と上記受信装置との間の通信環境に基づいて決定される通信システム。

２．　請求項１に記載の通信システムであって、

上記送信装置は、上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有する。

３．　直交周波数分割多重を利用して送信装置から第１の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

第１の受信装置へ伝送する第１の信号系列、および第１の受信装置とは異なる他の受信装置へ伝送する第２の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、

上記第１の信号系列の変調出力に第１のガード区間を、上記第２の信号系列の変調出力に第２のガード区間をそれぞれ挿入する挿入手段と、

上記第１のガード区間と第２のガード区間がそれぞれ挿入された変調信号

を送信する送信手段を有し、

上記第１の受信装置は、

上記第１のガード区間の削除処理と復調処理を行い、第１の信号系列を再生する復調手段を有し、

上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定されるとともに、上記第２のガード区間の長さは、上記送信装置と上記他の受信装置との間の通信環境に基づいて決定される。

４．　直交周波数分割多重を利用して送信装置から第１の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

第１の受信装置へ伝送する第１の信号系列、および上記送信装置の通信エリア内の第１の受信装置を含む複数の受信装置に伝送する第２の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、

上記第１の信号系列の変調出力に第１のガード区間を、上記第２の信号系列の変調出力に第２のガード区間をそれぞれ挿入する挿入手段と、

上記第１のガード区間と第２のガード区間がそれぞれ挿入された変調信号を送信する送信手段を有し、

上記第１の受信装置は、

上記第１のガード区間の削除処理と第２のガード区間の削除処理と復調処理を行い、第１の信号系列と第２の信号系列を再生する復調手段を有し、

上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定される

５．　請求項４に記載の通信システムであって、

上記第２ガード区間の長さは、通信エリア内に存在する複数の受信装置が第２の信号系列を再生できるように決定される。

６．　請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置と当該受信装置との間の回線の最大伝送遅延差を検出する検出手段をさらに有し、

上記挿入手段は、上記検出手段により検出された最大伝送遅延差に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その最大伝送遅延差に従ってガード区間を削除する。

７．　請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置と当該受信装置との間の伝送距離を推定する推定手段をさらに有し、

上記挿入手段は、上記推定手段により推定された伝送距離に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その推定された伝送距離に従ってガード区間を削除する。

８．　請求項１に記載の通信システムであって、

上記送信装置は、上記送信装置と当該受信装置との間の伝送距離を推定する推定手段をさらに有し、

上記挿入手段は、上記推定手段により推定された伝送距離に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その推定された伝送距離に従ってガード区間を削除する。

９．　請求項８に記載の通信システムであって、

上記推定手段は、当該送信装置から信号が送信されたときから、上記受信装置からその信号に対応する応答が返ってくるまでの時間に基づいて上記伝送距離を推定する。

１０．請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置から当該受信装置へ信号が伝送されたときの通信品質をモニタするモニタ手段をさらに有し、

上記ガード区間の長さは、予め決められた所定の通信品質が満たされるように決定される。

１１． 直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する方法であって、

直交周波数分割多重を利用して信号系列を変調し、

上記変調により得られた信号に対して、上記送信装置と上記受信装置との間の通信環境に基づいて決定される長さのガード区間を挿入し、

上記ガード区間が挿入された変調信号を送信する信号伝送方法。

１２． 請求項１１に記載の方法であって、

上記ガード区間が挿入された変調信号は、そのガード区間の長さに応じてその送信電力が制御される。

１３． 直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する方法であって、

送信相手先の異なる第１の信号系列および第２の信号系列を直交周波数多重を利用して変調し、

上記変調された第１の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第１のガード区間を挿入し、上記変調された第２の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第２のガード区間を挿入し、

上記第１のガード区間および第２のガード区間がそれぞれ挿入された変調信号を送信する信号伝送方法。

１４． 請求項１１に記載の方法であって、

上記ガード区間の長さは、上記送信装置と上記受信装置との間の回線の最大伝送遅延差または伝送距離に基づいて決定される。

１５． 請求項１１に記載の方法であって、

上記送信装置から上記受信装置へ信号が伝送されたときの通信品質をモニタし、

上記ガード区間の長さは、予め決められた所定の通信品質が満たされるように決定される。

１６． セルラ通信システムにおいて直交周波数分割多重を利用して移動機へ信号を伝送する基地局装置であって、

直交周波数分割多重を利用して信号系列を変調する変調手段と、

上記変調手段により得られた変調信号に対して、当該基地局装置と上記信号系列を送信すべき移動機との間の通信環境に基づいて決定される長さのガード区間を挿入する挿入手段と、

上記ガード区間が挿入された変調信号を送信する送信手段と、

を有する基地局装置。

１７． 請求項１６に記載の基地局装置であって、

上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有する。

１８． セルラ通信システムにおいて直交周波数分割多重を利用して移動機へ信号を伝送する基地局装置であって、

送信相手先の異なる第１の信号系列および第２の信号系列を直交周波数多重を利用してそれぞれ変調する変調手段と、

上記変調手段により得られた変調された第１の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第１のガード区間を挿入するとともに、上記変調手段により得られた変調された第２の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第２のガード区間を挿入する挿入手段と、

上記第１のガード区間および第２のガード区間がそれぞれ挿入された変調信

号を送信する送信手段と、

を有する基地局装置。

１９．　セルラ通信システムにおいて、直交周波数分割多重を利用して基地局から送信された信号を受信する移動機であって、

受信した信号が自移動機宛ての第１の信号系列と他移動機宛ての第２の信号系列を含む場合、第１の信号系列に対応する第１のガード区間の削除処理と復調処理を行う復調手段を有する移動機。

２０．　セルラ通信システムにおいて、直交周波数分割多重を利用して基地局から送信された信号を受信する移動機であって、

受信した信号が自移動機宛ての第１の信号系列と自移動機を含む複数の移動機宛ての第２の信号系列を含む場合、第１の信号系列に対応する第１のガード区間の削除処理と、第２の信号系列に対応する第２のガード区間の削除処理と、復調処理を行う復調手段を有する移動機。

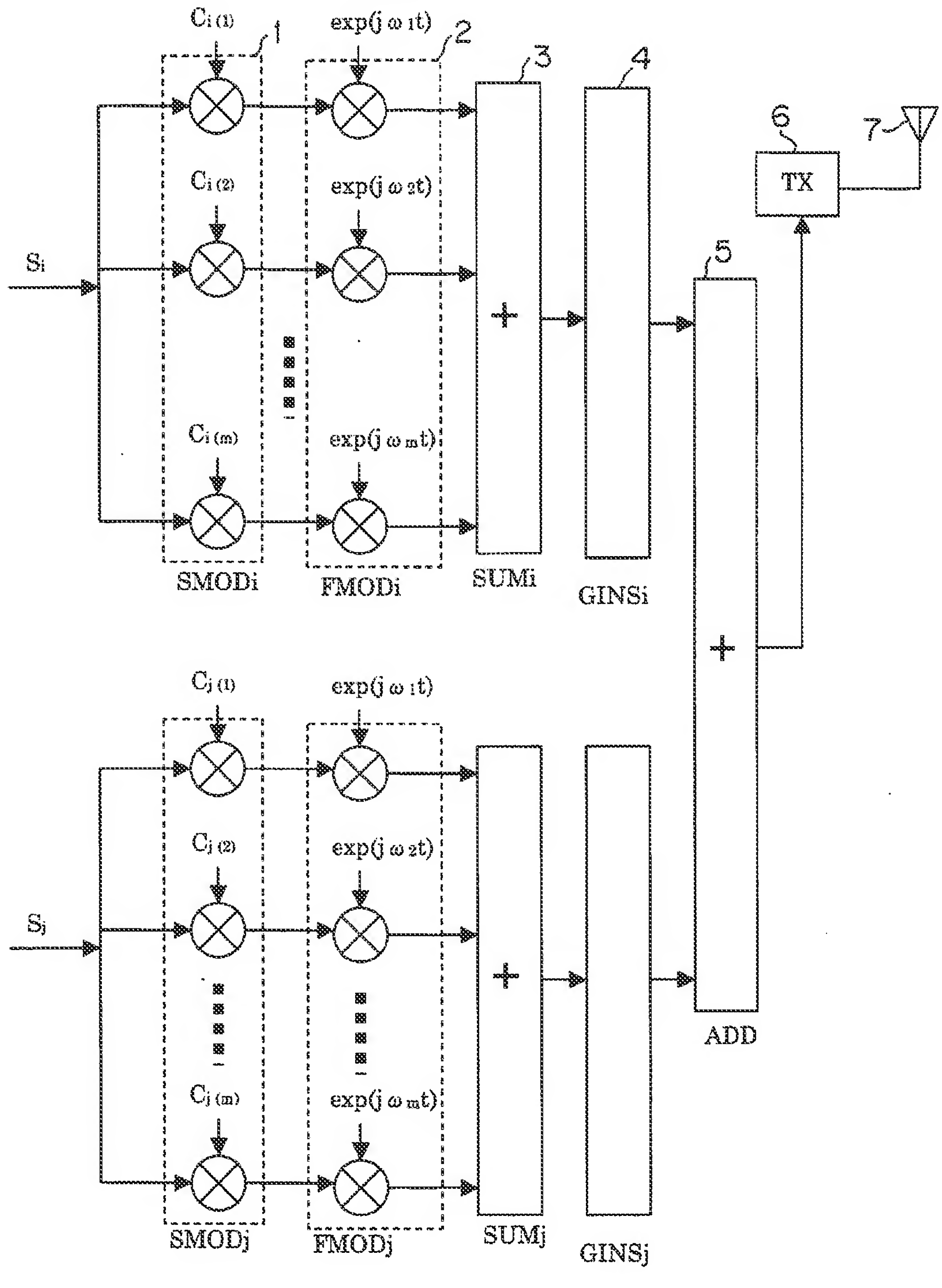

Si
Ci(1)
Ci(2)
Ci(m)
exp(jω1t)
exp(jω2t)
exp(jωmt)
1
2
3
4
5
6
7
TX
+
SMODi
FMODi
SUMi
GINSi
ADD
Sj
Cj(1)
Cj(2)
Cj(m)
SMODj
FMODj
SUMj
GINSj

図1

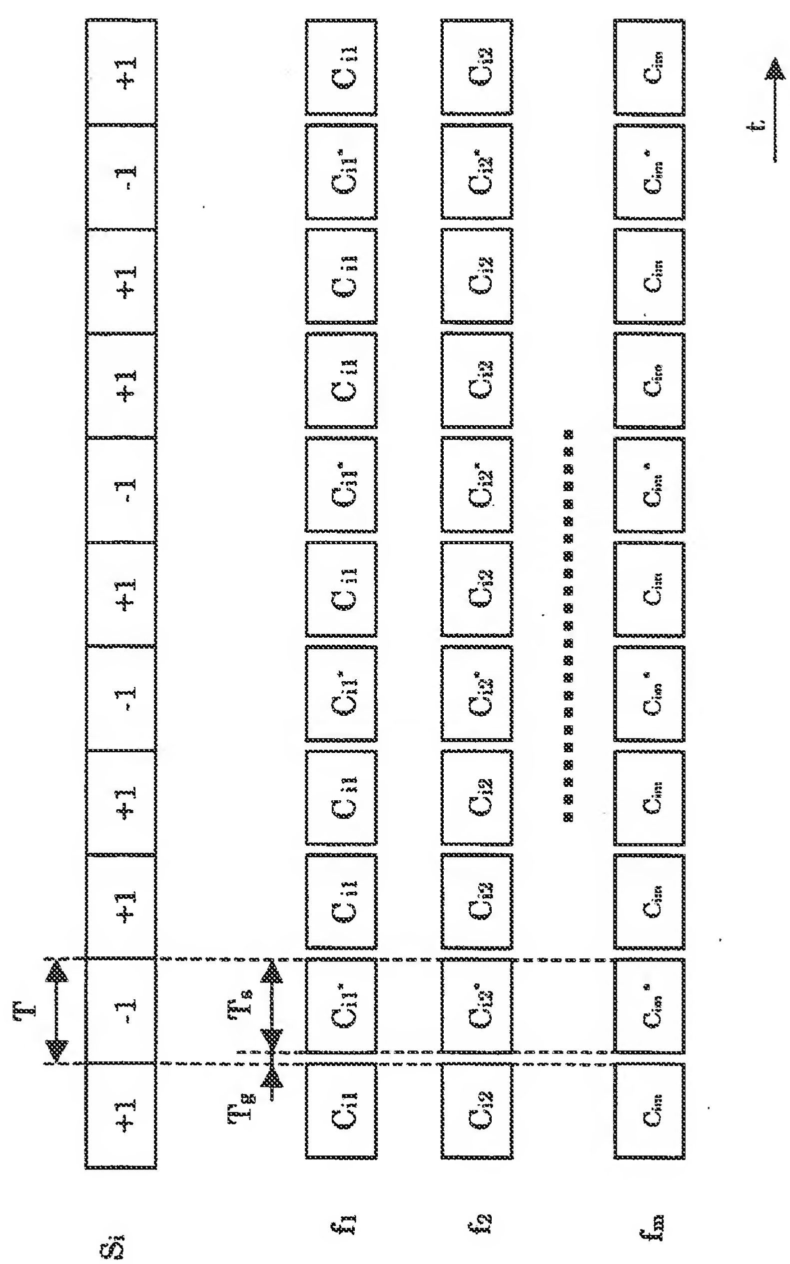


図2

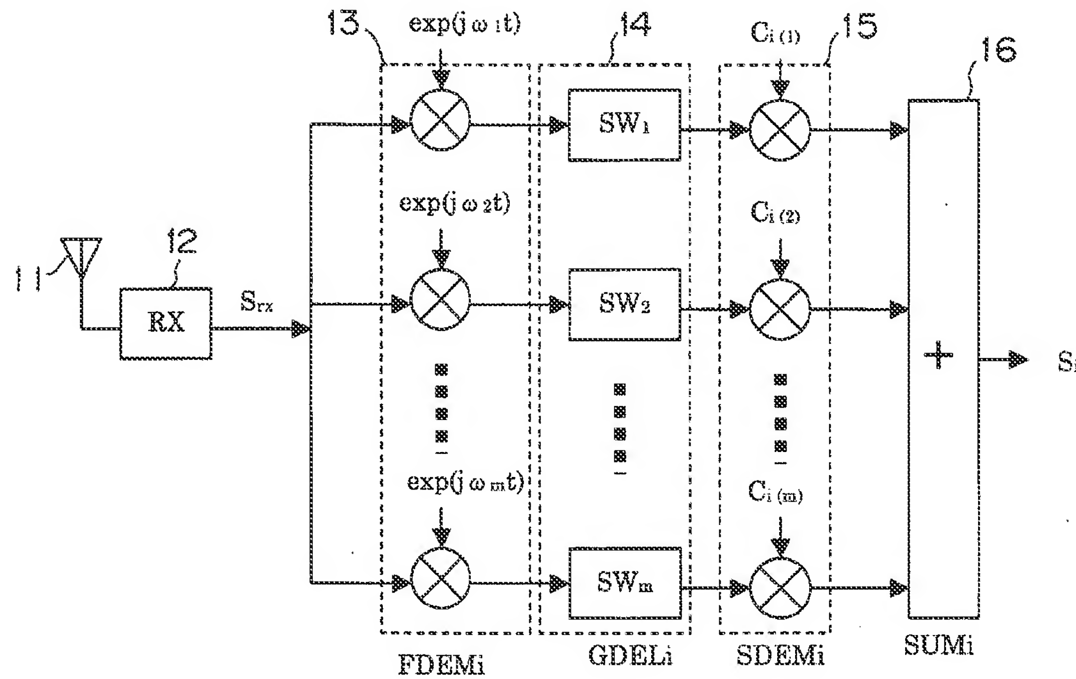

13
exp(jω1t)
14
Ci(1)
15
16
11
12
RX
Srx
exp(jω2t)
exp(jωmt)
SW1
SW2
SWm
Ci(2)
Ci(m)
+
Si
FDEMi
GDELi
SDEMi
SUMi

図3

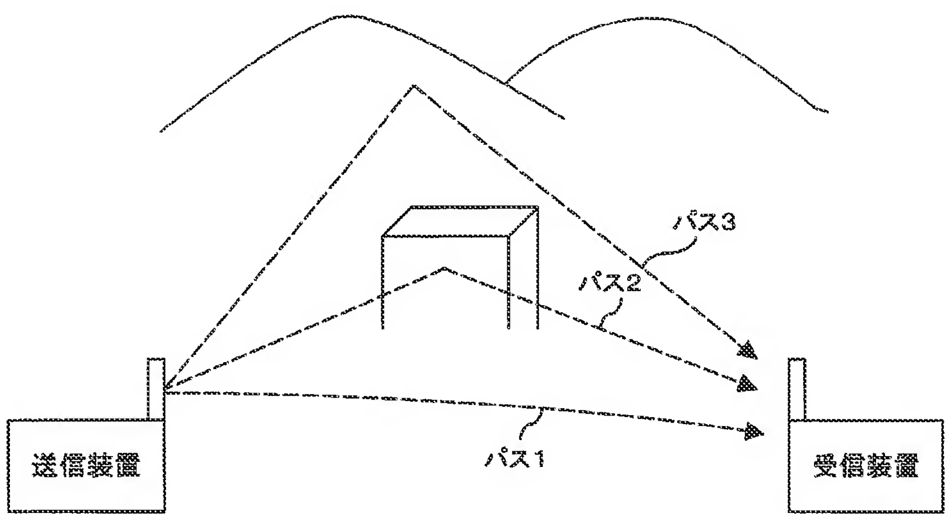


図4

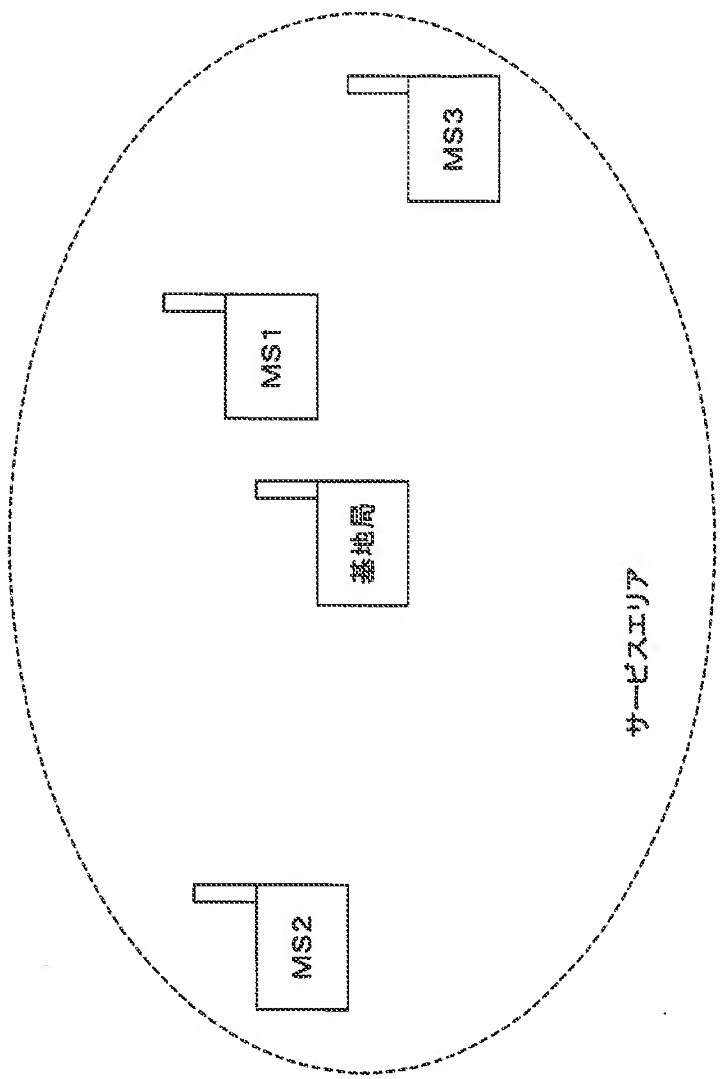


図5

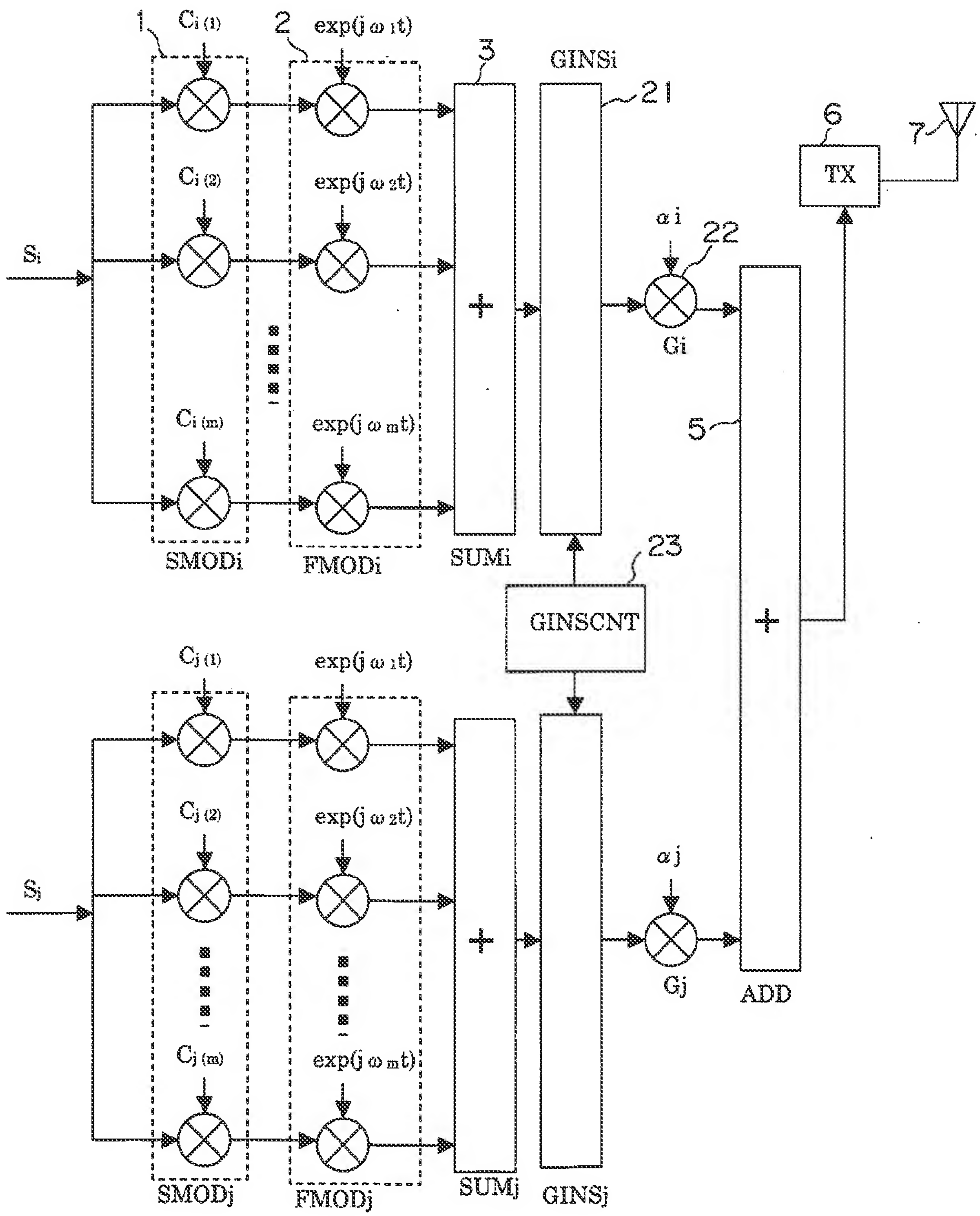

1
2
3
6
7
21
22
23
5
$C_{i(1)}$
$C_{i(2)}$
$C_{i(m)}$
$\exp(j\omega_1 t)$
$\exp(j\omega_2 t)$
$\exp(j\omega_m t)$
S_i
GINSi
TX
αi
Gi
SMODi
FMODi
SUMi
GINSCNT
$C_{j(1)}$
$C_{j(2)}$
$C_{j(m)}$
S_j
αj
Gj
ADD
SMODj
FMODj
SUMj
GINSj

図6

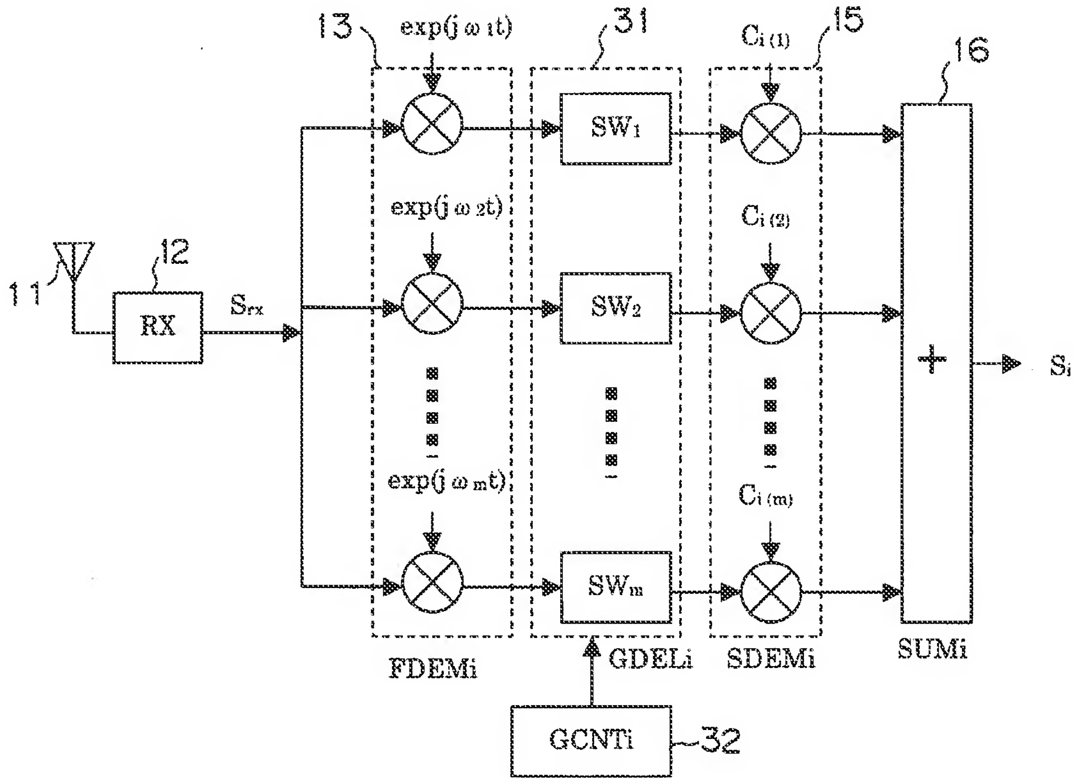

13
exp(j ω 1t)
31
15
Ci (1)
16
SW1
exp(j ω 2t)
Ci (2)
12
11
RX
Srx
SW2
+
Si
exp(j ω mt)
Ci (m)
SWm
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図7

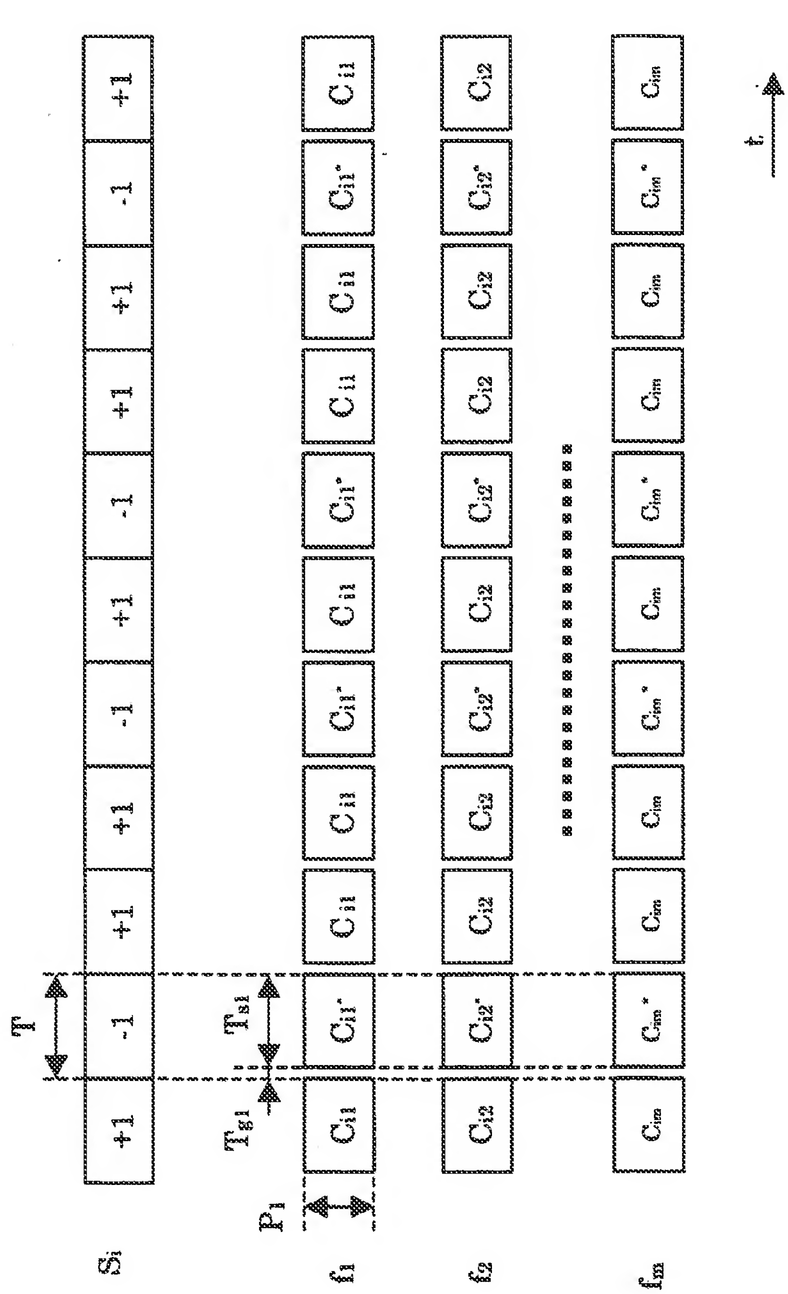


図8

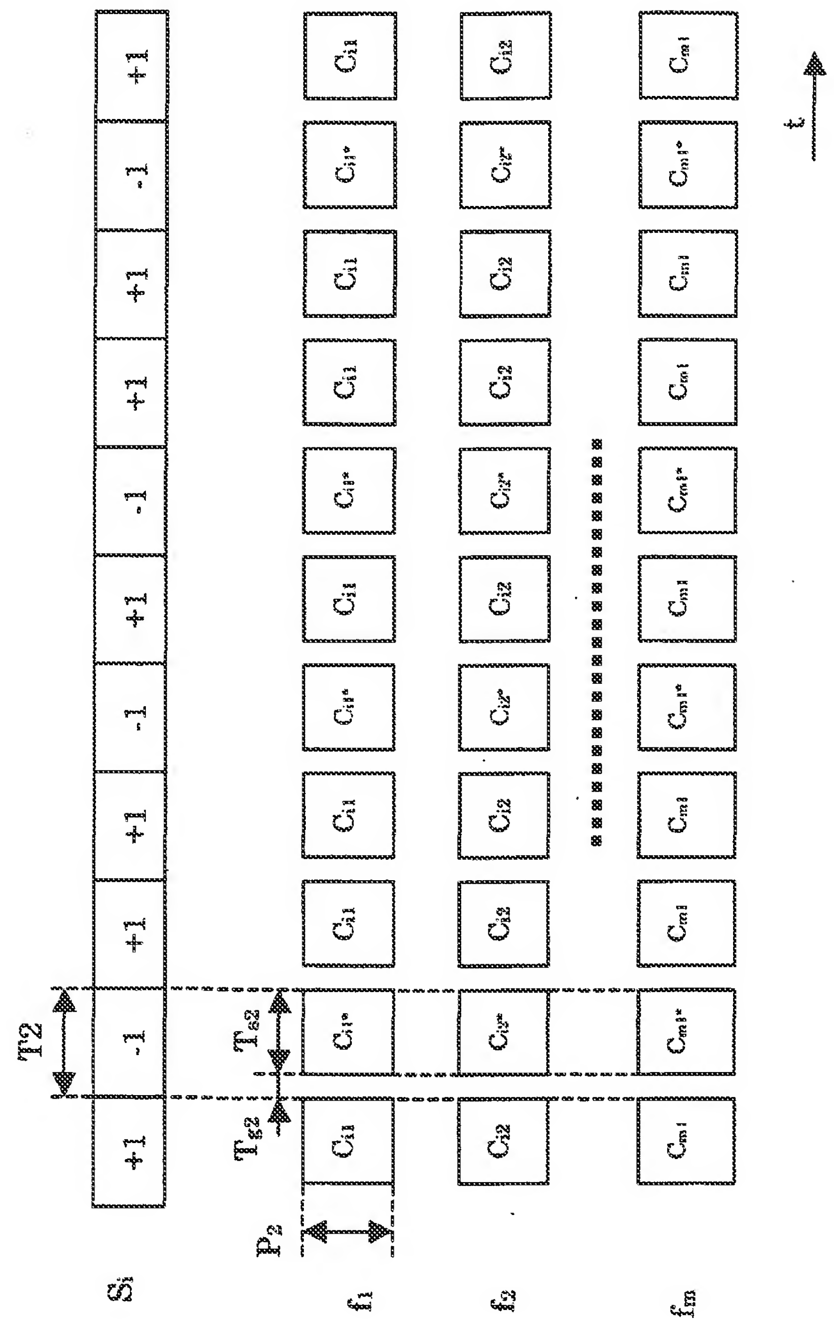


図9

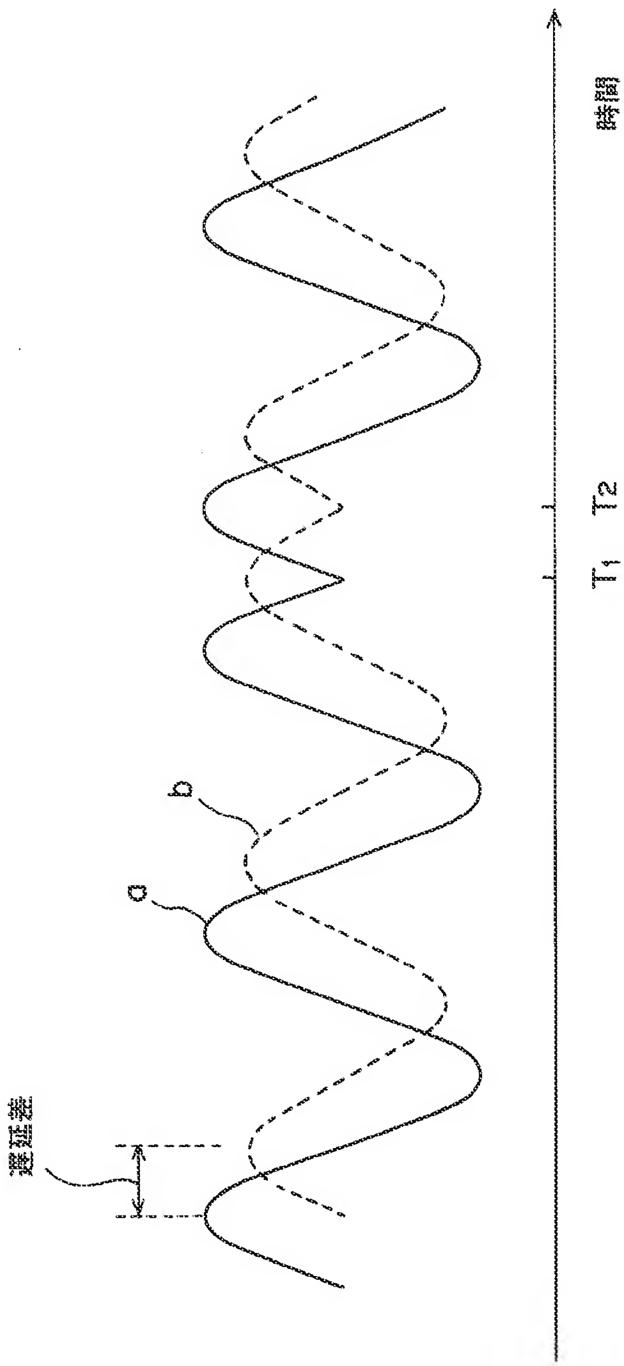


図10

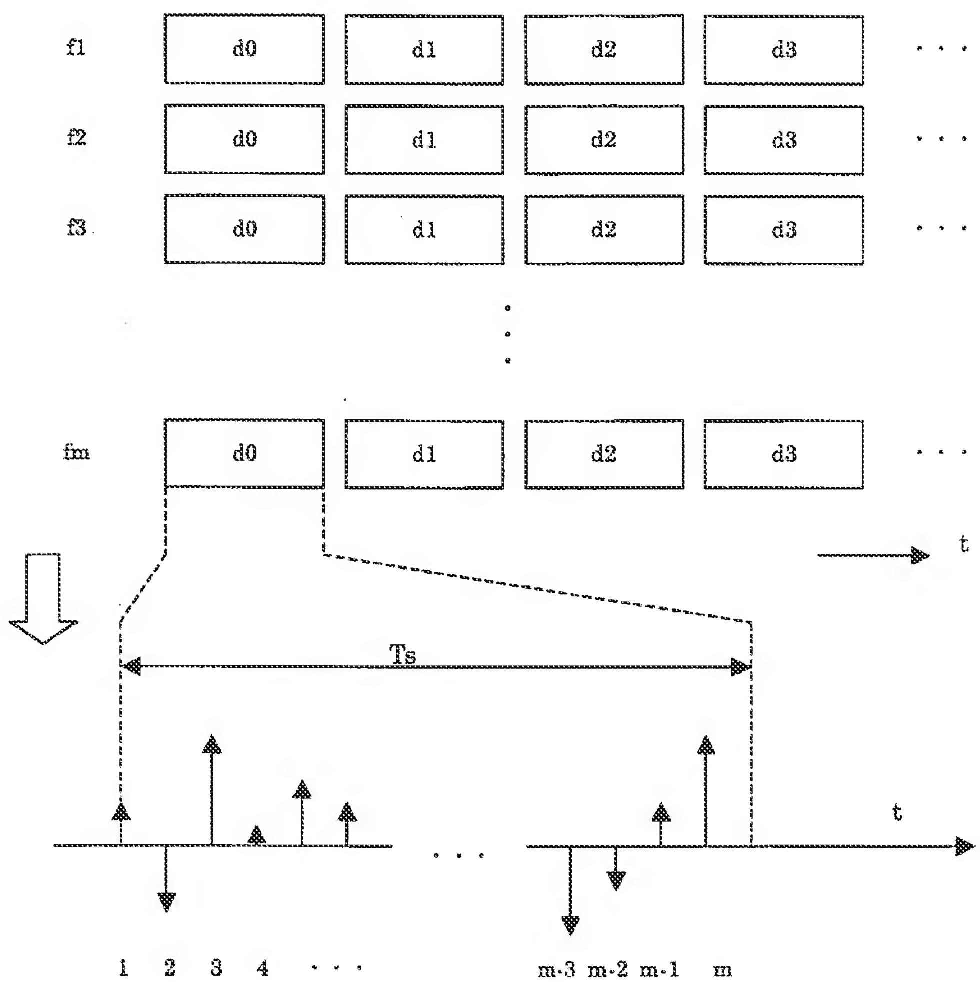

f1
d0
d1
d2
d3
f2
d0
d1
d2
d3
f3
d0
d1
d2
d3
fm
d0
d1
d2
d3
t
Ts
t
1 2 3 4
m-3 m-2 m-1 m

図11

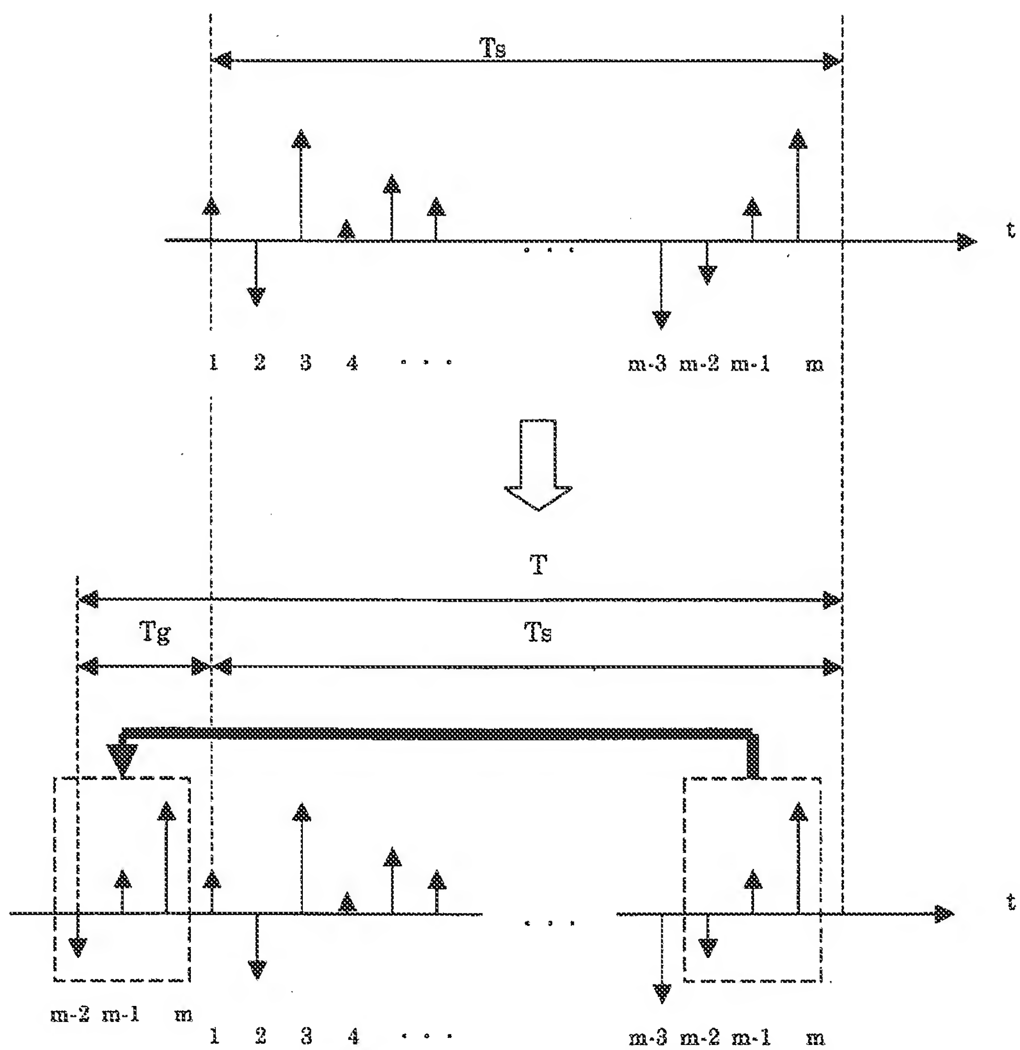


図12

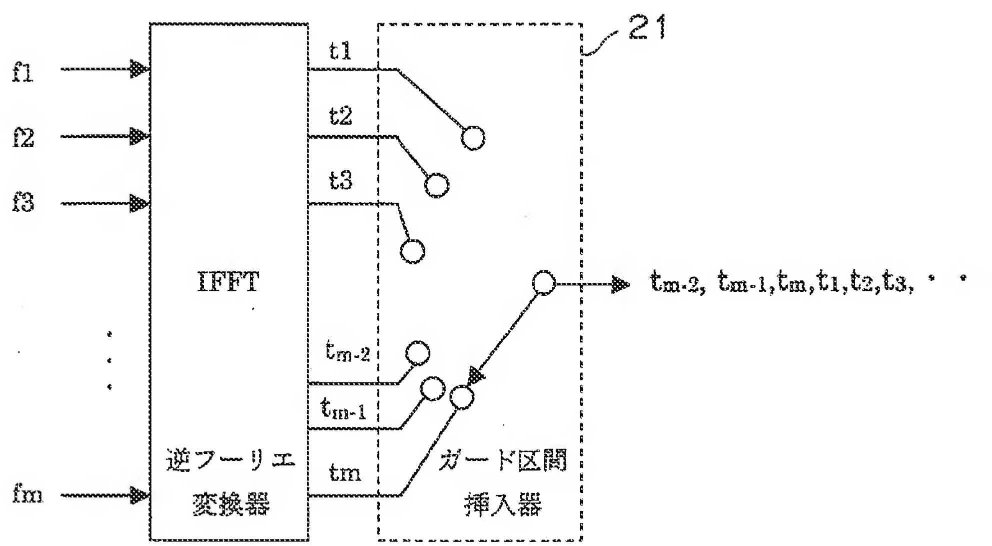

21
f1
f2
f3
fm
IFFT
逆フーリエ
変換器
t1
t2
t3
t_{m-2}
t_{m-1}
tm
ガード区間
挿入器
t_{m-2}, t_{m-1}, t_m, t_1, t_2, t_3, ・・・

図13

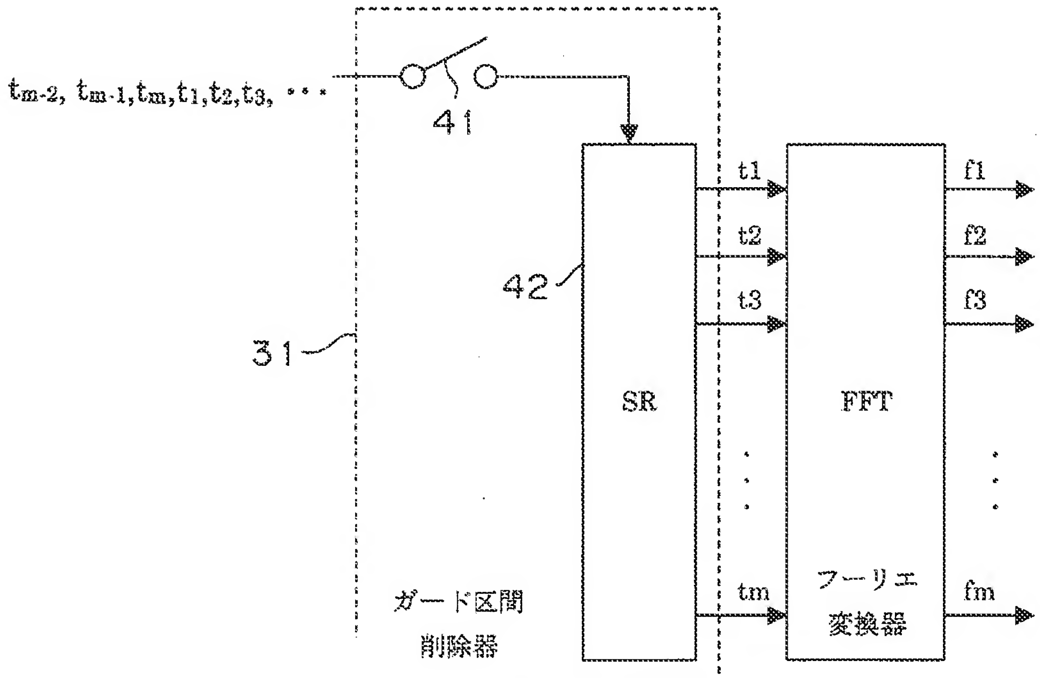

t_{m-2}, t_{m-1},t_m,t_1,t_2,t_3, ・・・
41
42
31
SR
FFT
t1
t2
t3
tm
f1
f2
f3
fm
ガード区間
削除器
フーリエ
変換器

図14

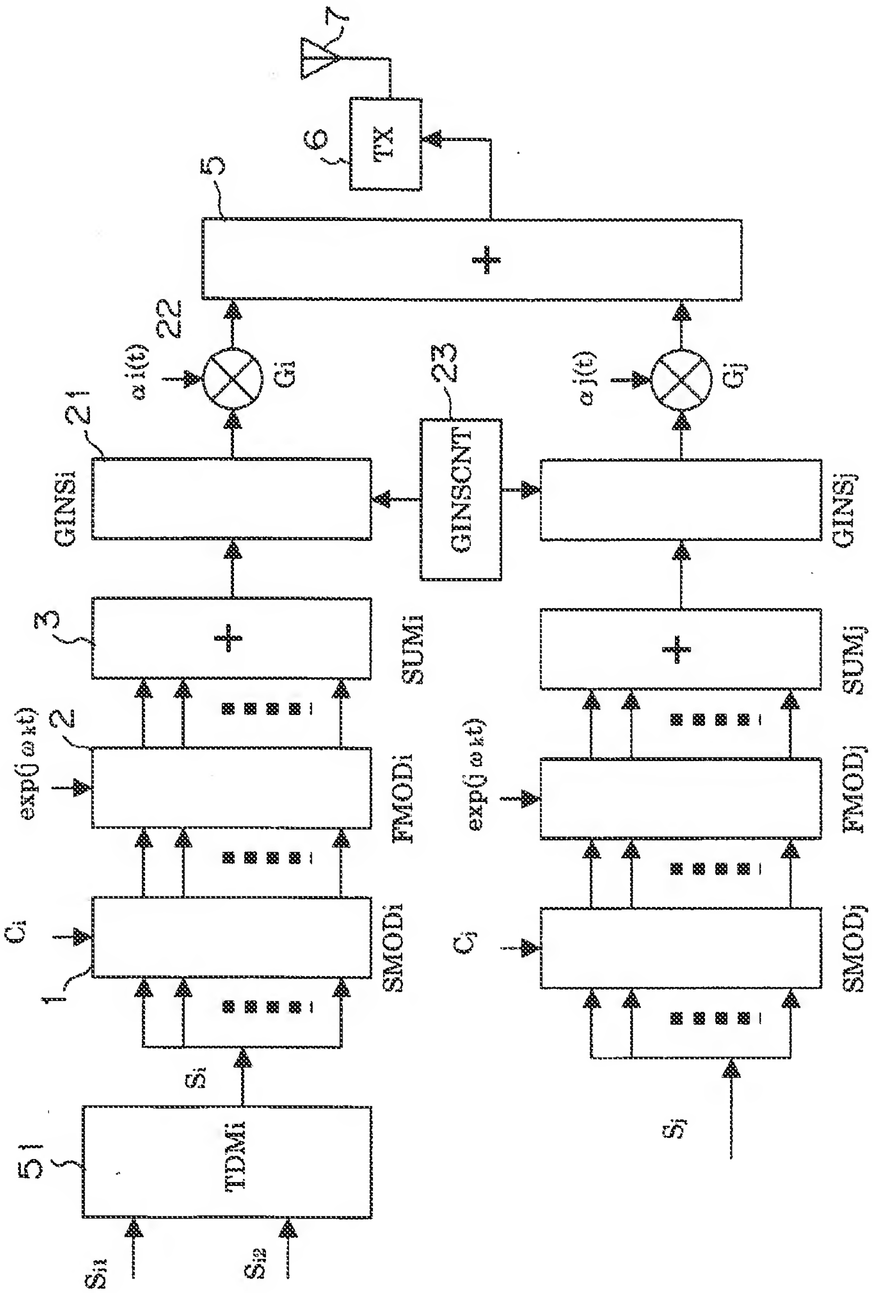

7
TX
6
5
+
22
αi(t)
Gi
21
GINSi
23
GINSCNT
αj(t)
Gj
GINSj
3
+
SUMi
+
SUMj
2
exp(jωkt)
FMODi
exp(jωkt)
FMODj
Ci
SMODi
1
Cj
SMODj
Si
Sj
51
TDMi
Si1
Si2

図15

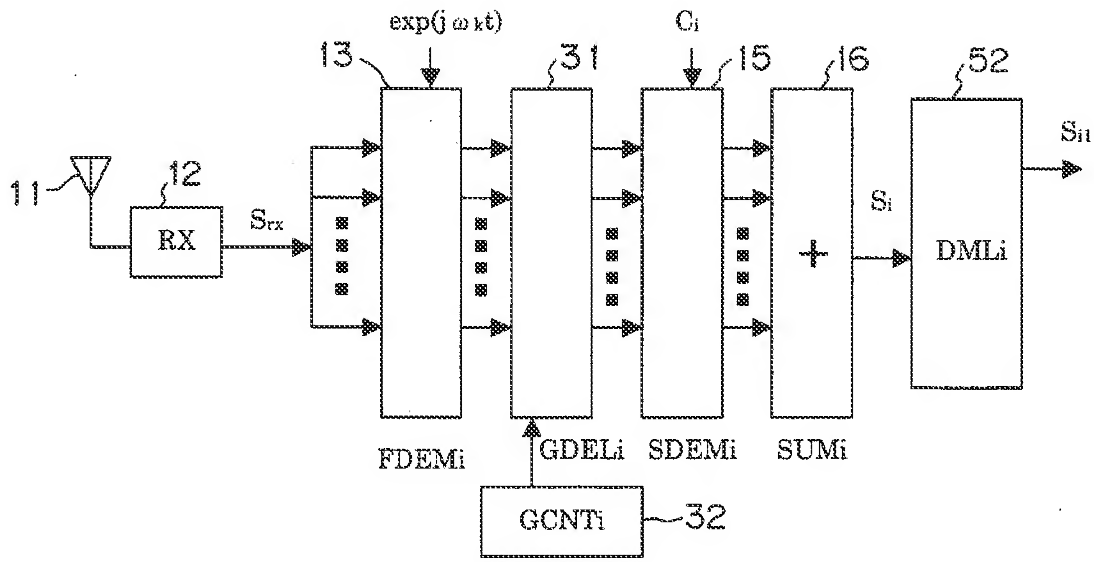
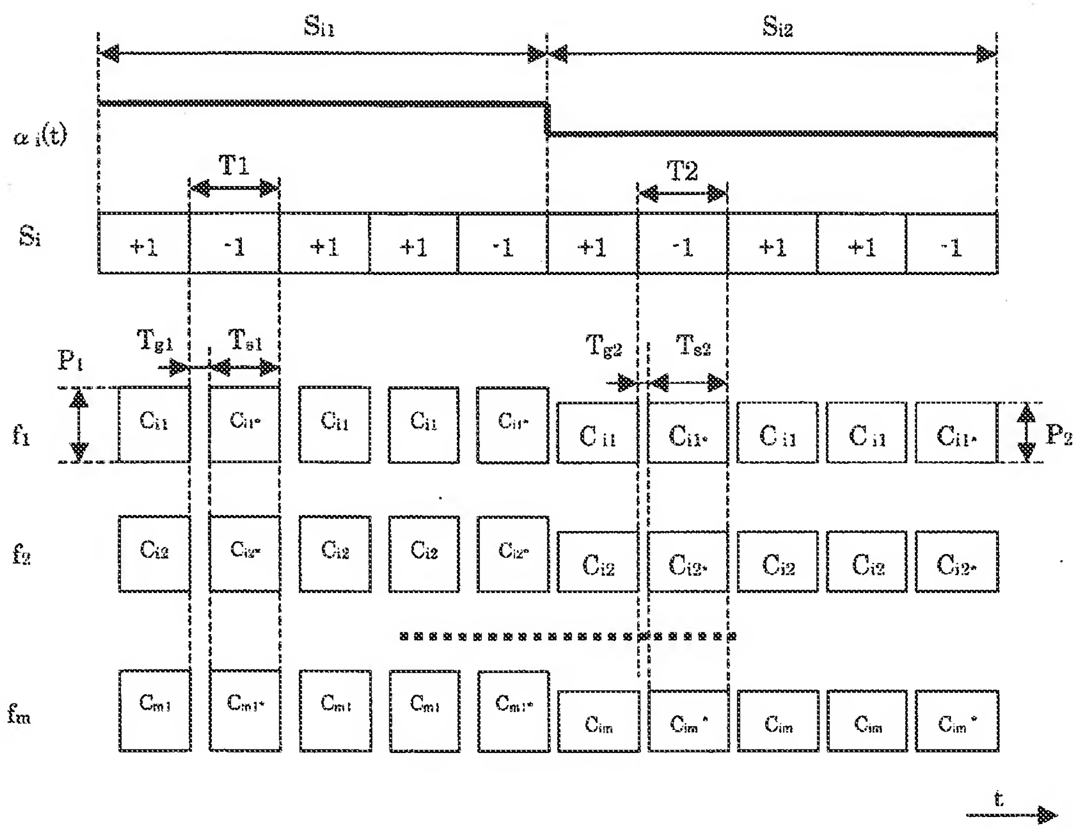

exp(jωkt)
Ci
13
31
15
16
52
11
12
RX
Srx
Si
Si1
+
DMLi
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図16



S_{i1}
S_{i2}
$\alpha_i(t)$
T1
T2
S_i
+1
-1
+1
+1
-1
+1
-1
+1
+1
-1
T_{g1}
T_{s1}
T_{g2}
T_{s2}
P_1
P_2
f_1
f_2
f_m
t

図17

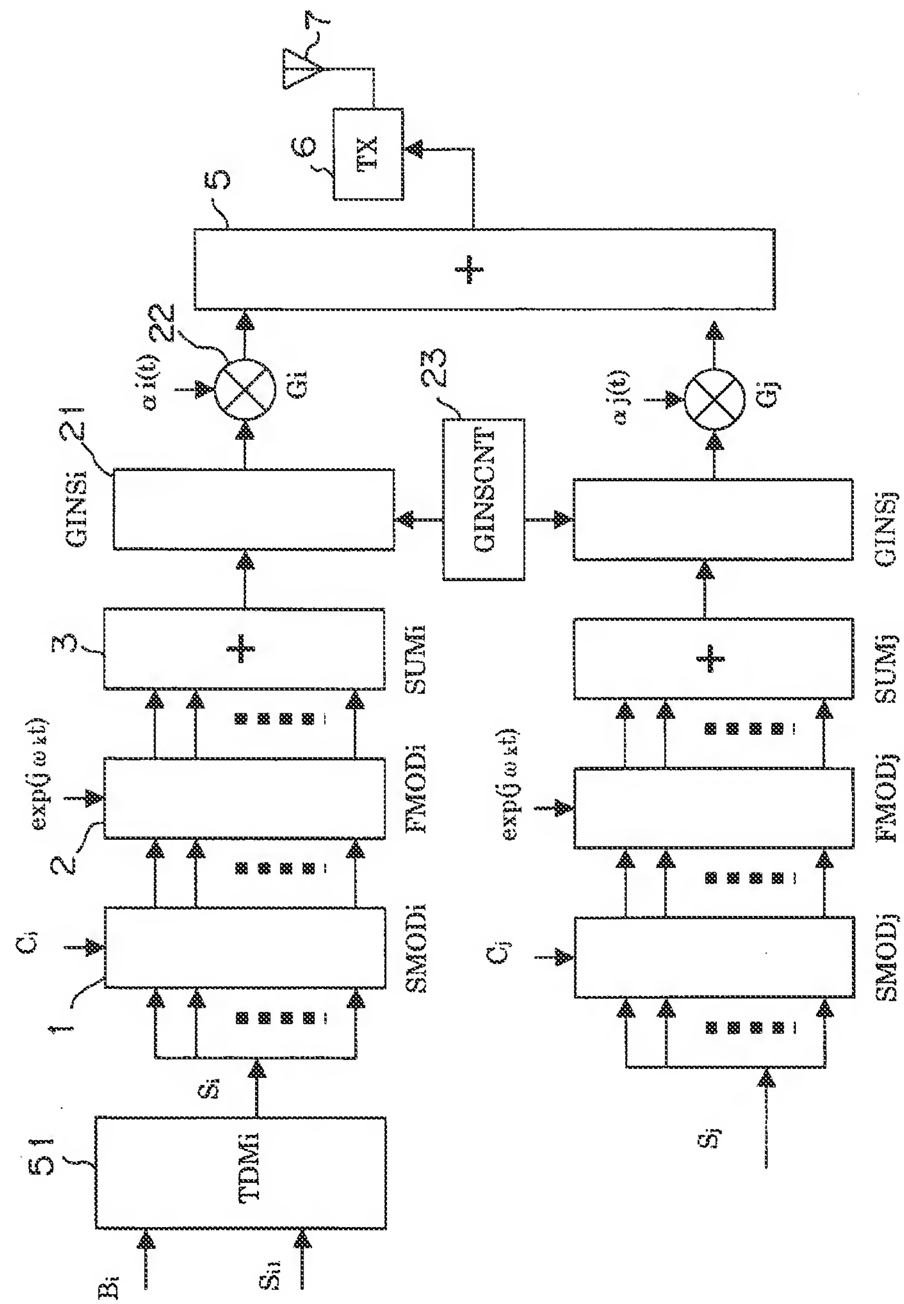

7
TX
6
5
+
22
αi(t)
Gi
23
GINSCNT
αj(t)
Gj
21
GINSi
GINSj
3
+
SUMi
SUMj
exp(jωkt)
FMODi
FMODj
2
Ci
Cj
SMODi
SMODj
1
Si
Sj
51
TDMi
Bi
Si1

図18

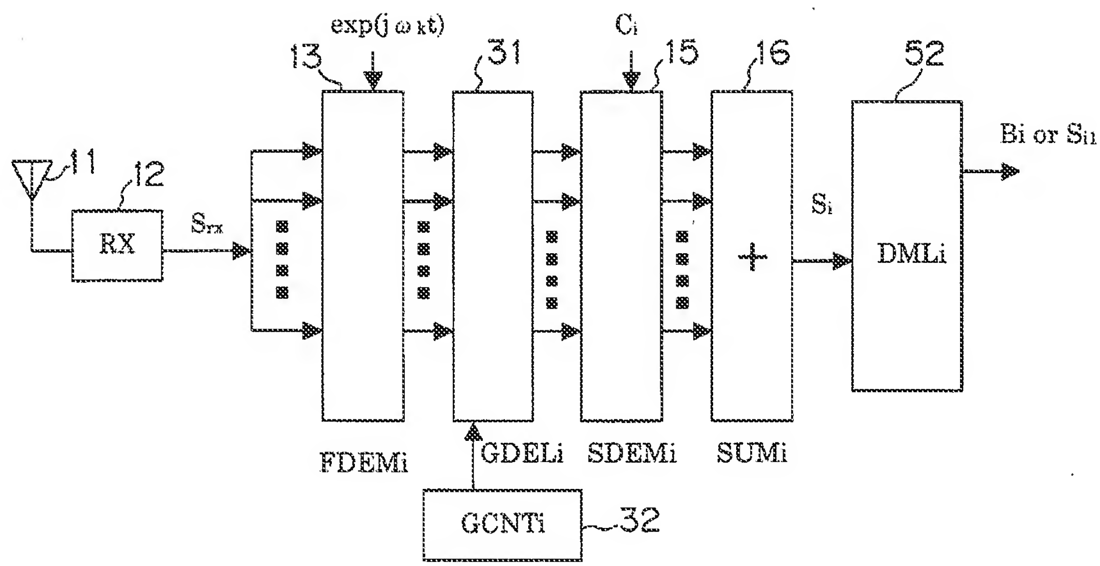

exp(jωkt)
13
31
Ci
15
16
52
11
12
RX
Srx
+
Si
DMLi
Bi or Si1
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図19

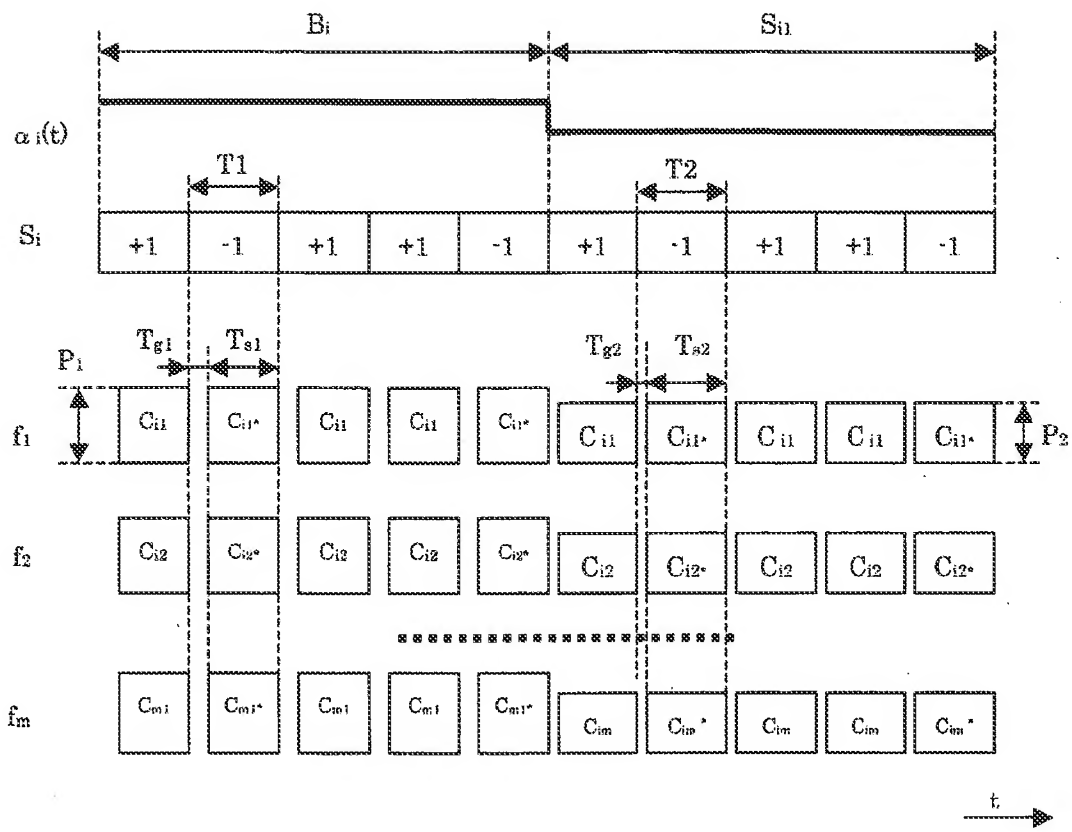
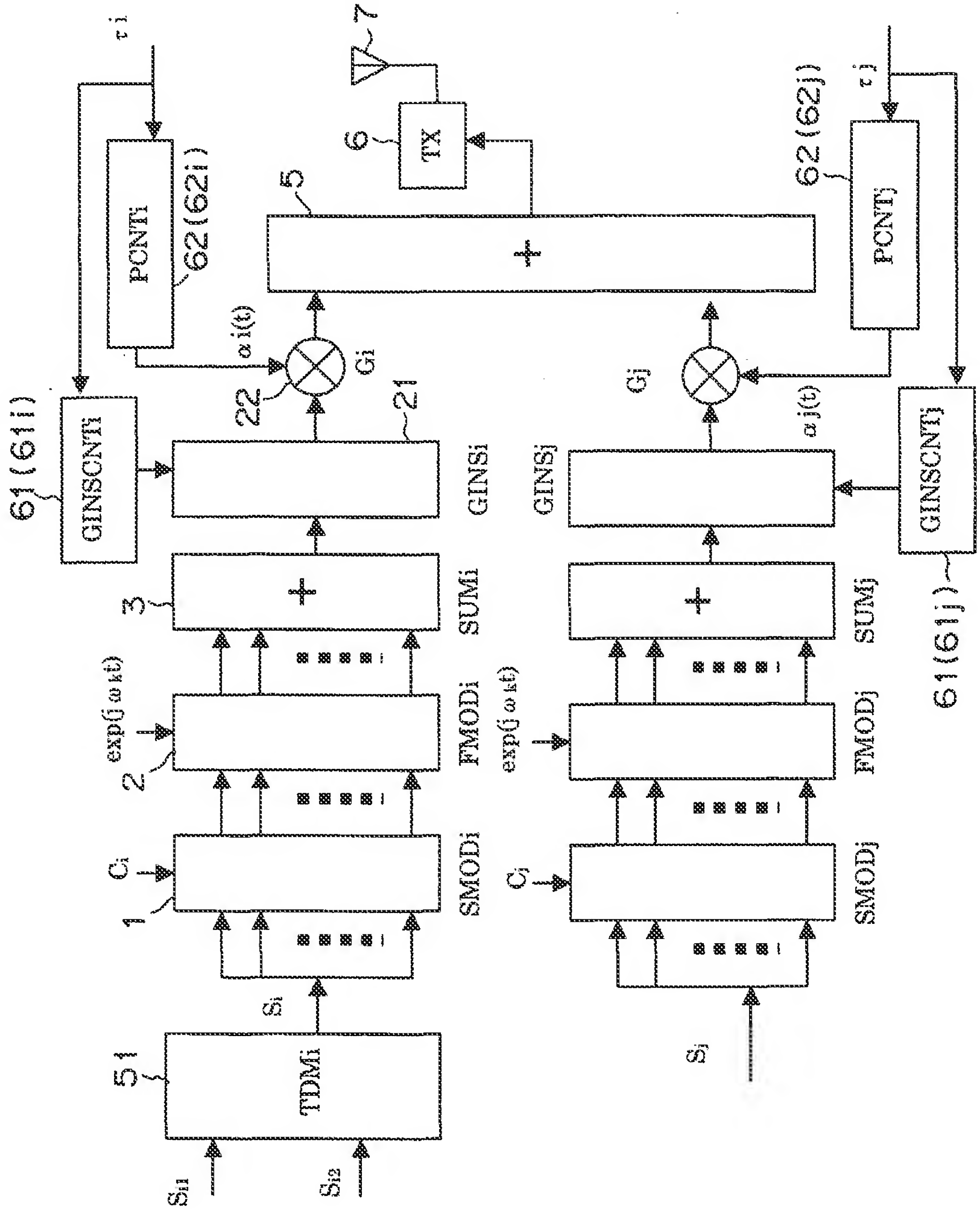


図20



TDMi
51
S_{i1}
S_{i2}
S_i
SMODi
1
C_i
FMODi
2
exp(jωkt)
SUMi
3
GINSi
21
GINSCNTi
61(61i)
22
Gi
αi(t)
PCNTi
62(62i)
τi
5
TX
6
7
S_j
SMODj
C_j
FMODj
exp(jωkt)
SUMj
GINSj
GINSCNTj
61(61j)
Gj
αj(t)
PCNTj
62(62j)
τj

図21

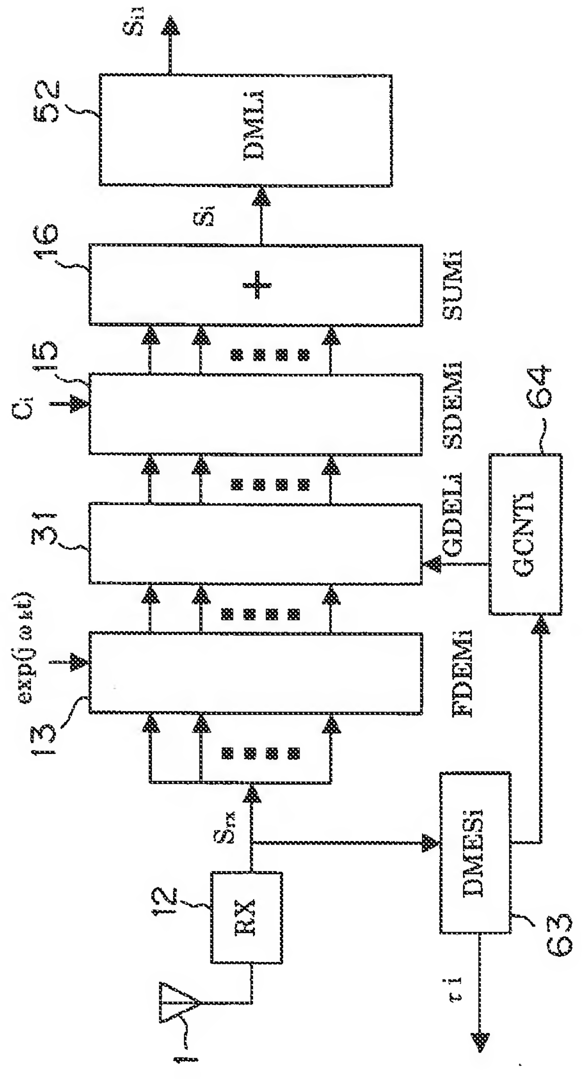

11
12
RX
S_{rx}
13
$exp(j\omega_k t)$
FDEMi
31
GDELi
15
C_i
SDEMi
16
SUMi
S_i
52
DMLi
S_{i1}
63
DMESi
τi
64
GCNTi

図22

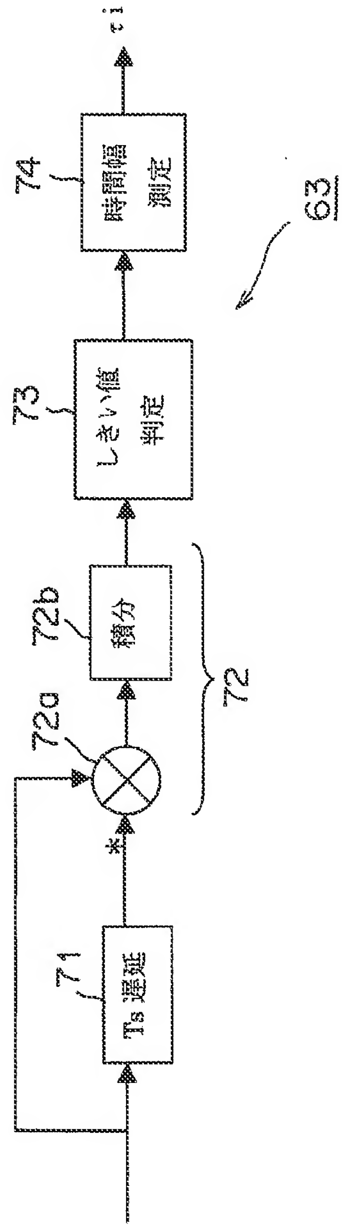


図23

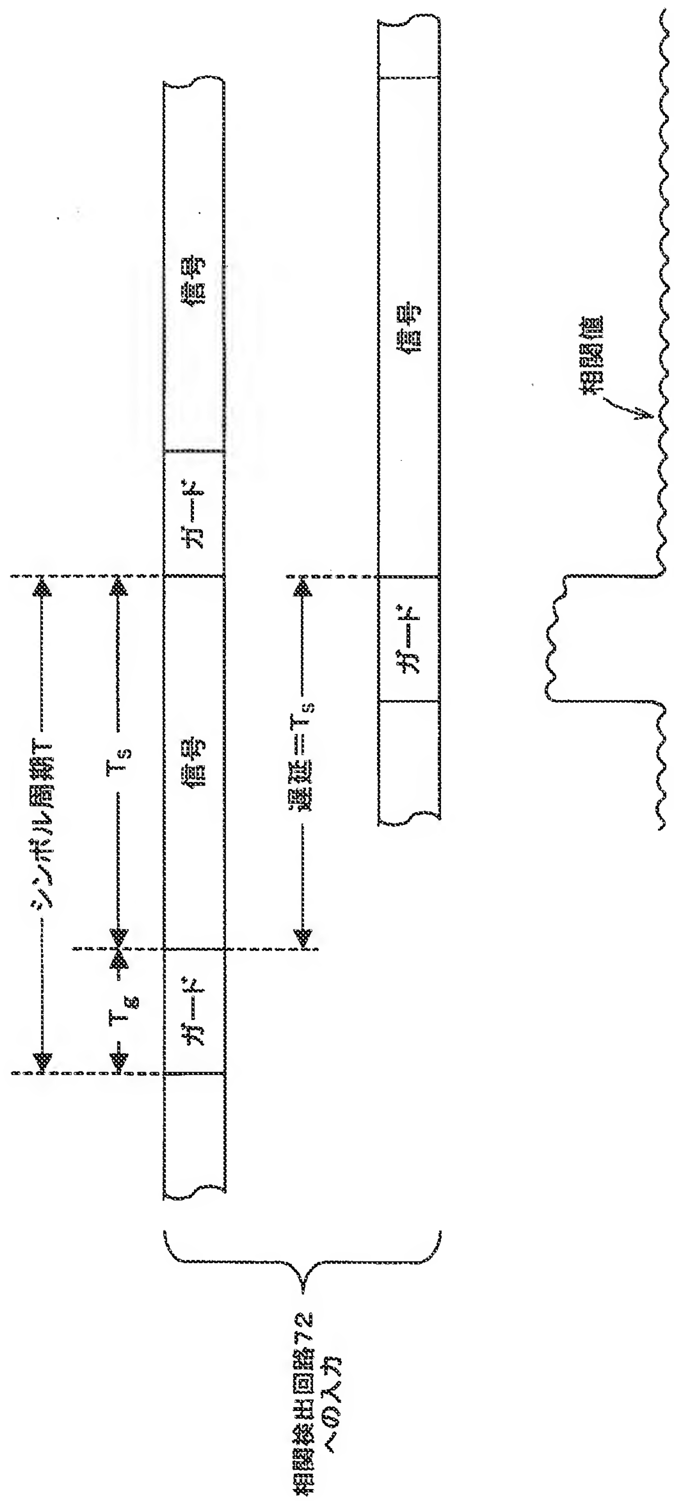


図24

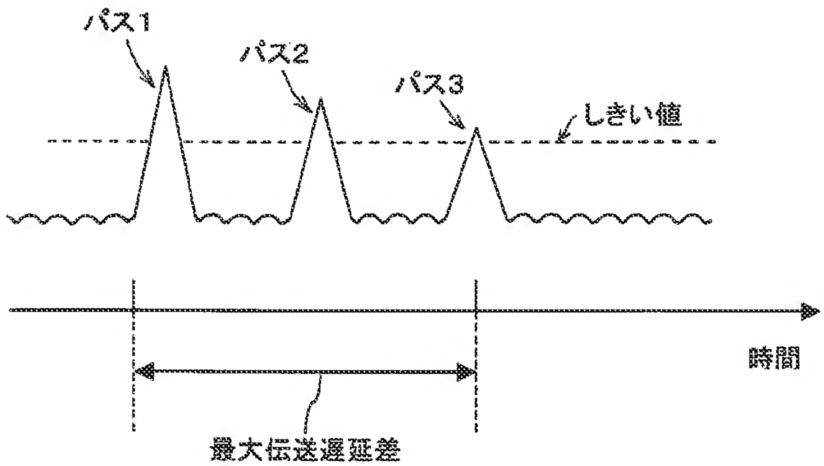


図25

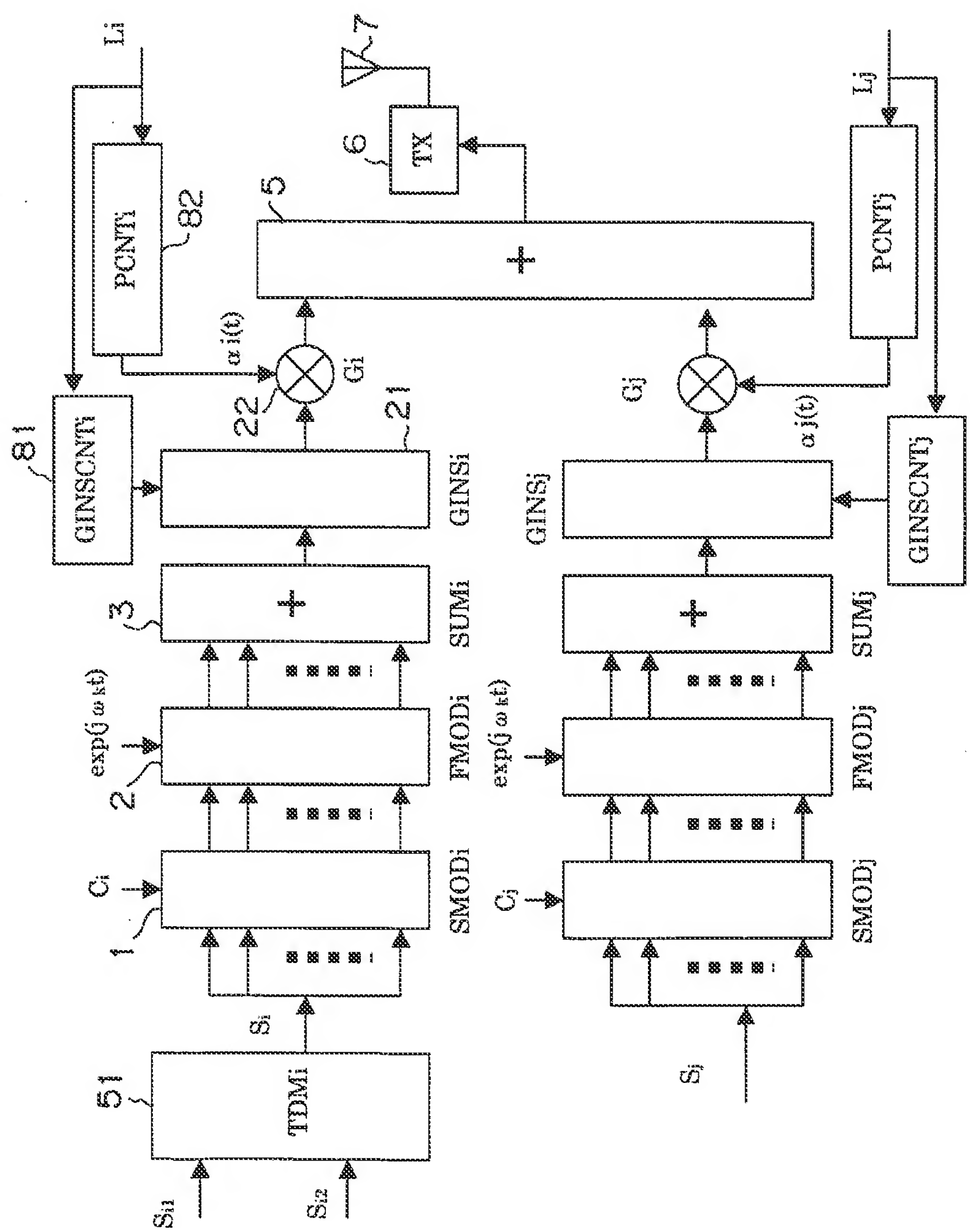
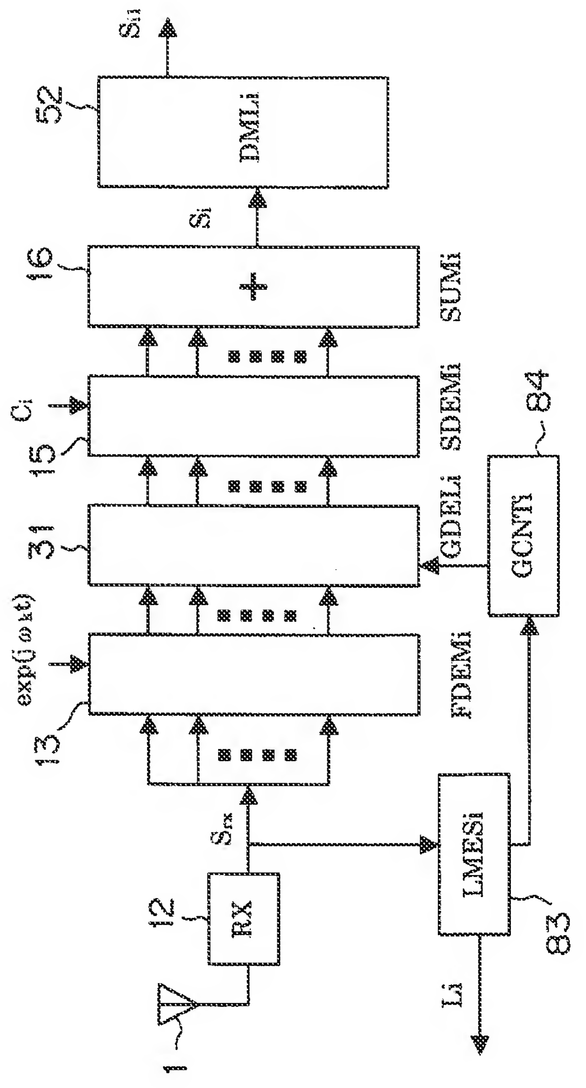

Li
Lj
7
6
TX
5
+
82
PCNTi
PCNTj
81
GINSCNTi
GINSCNTj
αi(t)
αj(t)
22
Gi
Gj
21
GINSi
GINSj
3
SUMi
SUMj
2
exp(jωkt)
FMODi
FMODj
1
Ci
Cj
SMODi
SMODj
Si
Sj
51
TDMi
Si1
Si2

図26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12
RX
Srx
13
exp(jωkt)
FDEMi
31
GDELi
15
Ci
SDEMi
16
+
SUMi
Si
52
DMLi
Si1
LMESi
83
Li
GCNTi
84

図27

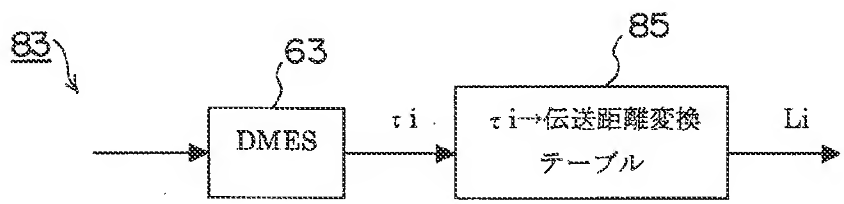


図28

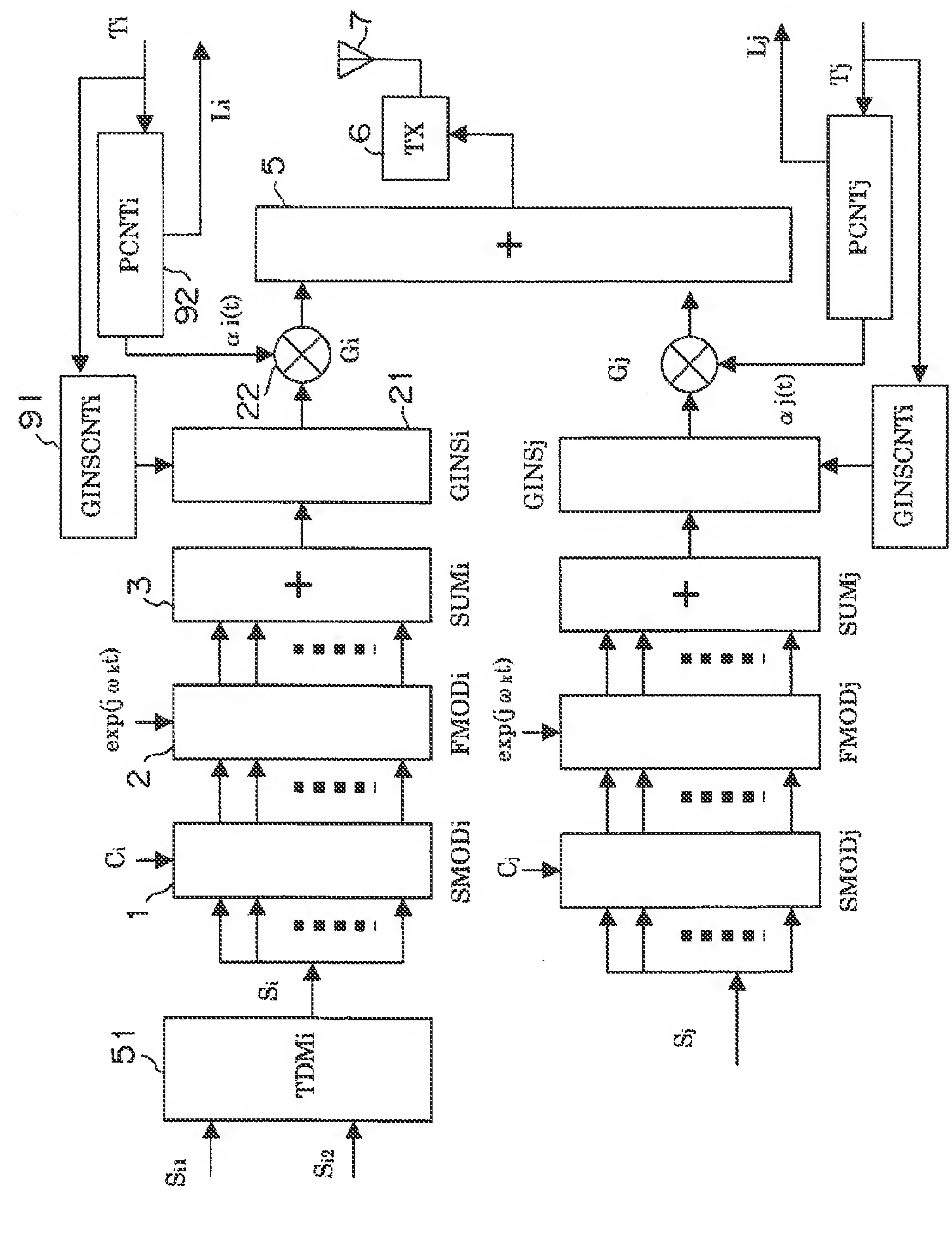
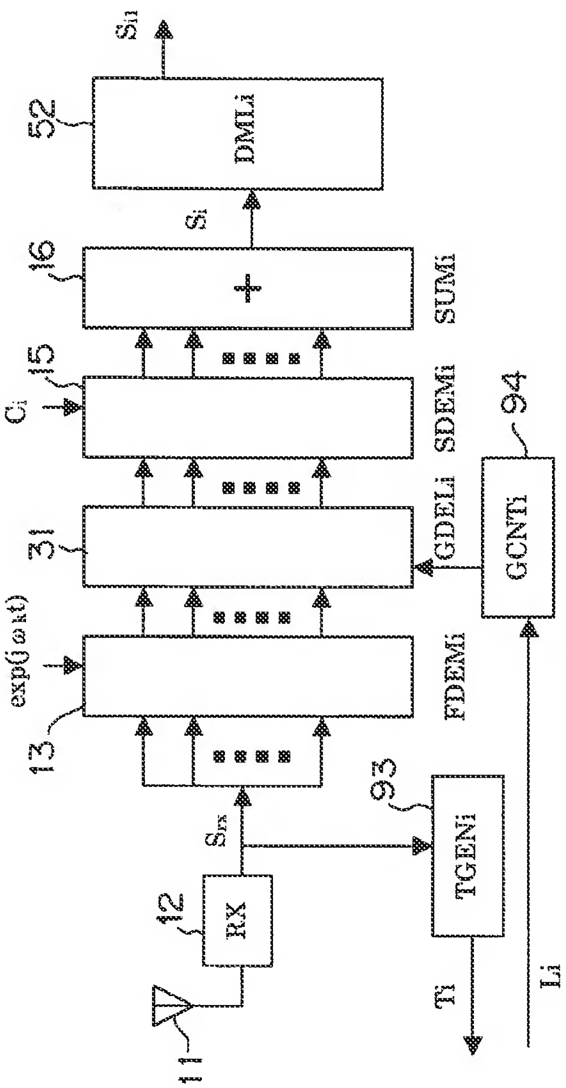

TDMi
SMODi
FMODi
SUMi
GINSi
GINSCNTi
PCNTi
TX
SMODj
FMODj
SUMj
GINSj
GINSCNTj
PCNTj
Gi
Gj
Li
Lj
Ti
Tj
exp(jωkt)
α i(t)
α j(t)
51
1
2
3
21
22
91
92
5
6
7

図29


11
12
RX
Srx
13
exp(jωkt)
FDEMi
31
GDELi
15
Ci
SDEMi
16
+
SUMi
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DMLi
Si1
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TGENi
Ti
94
GCNTi
Li

図30

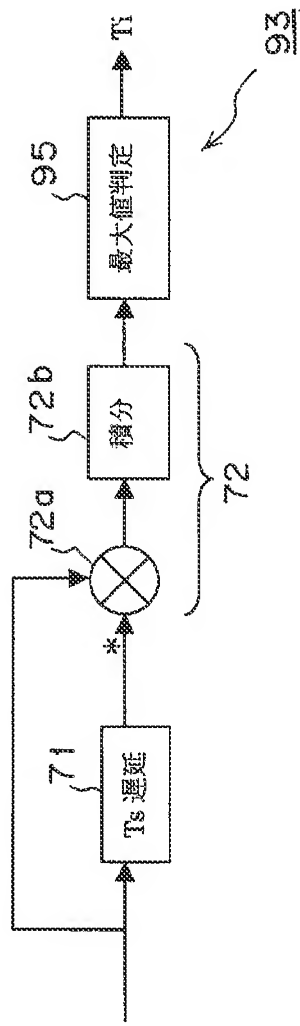


図31

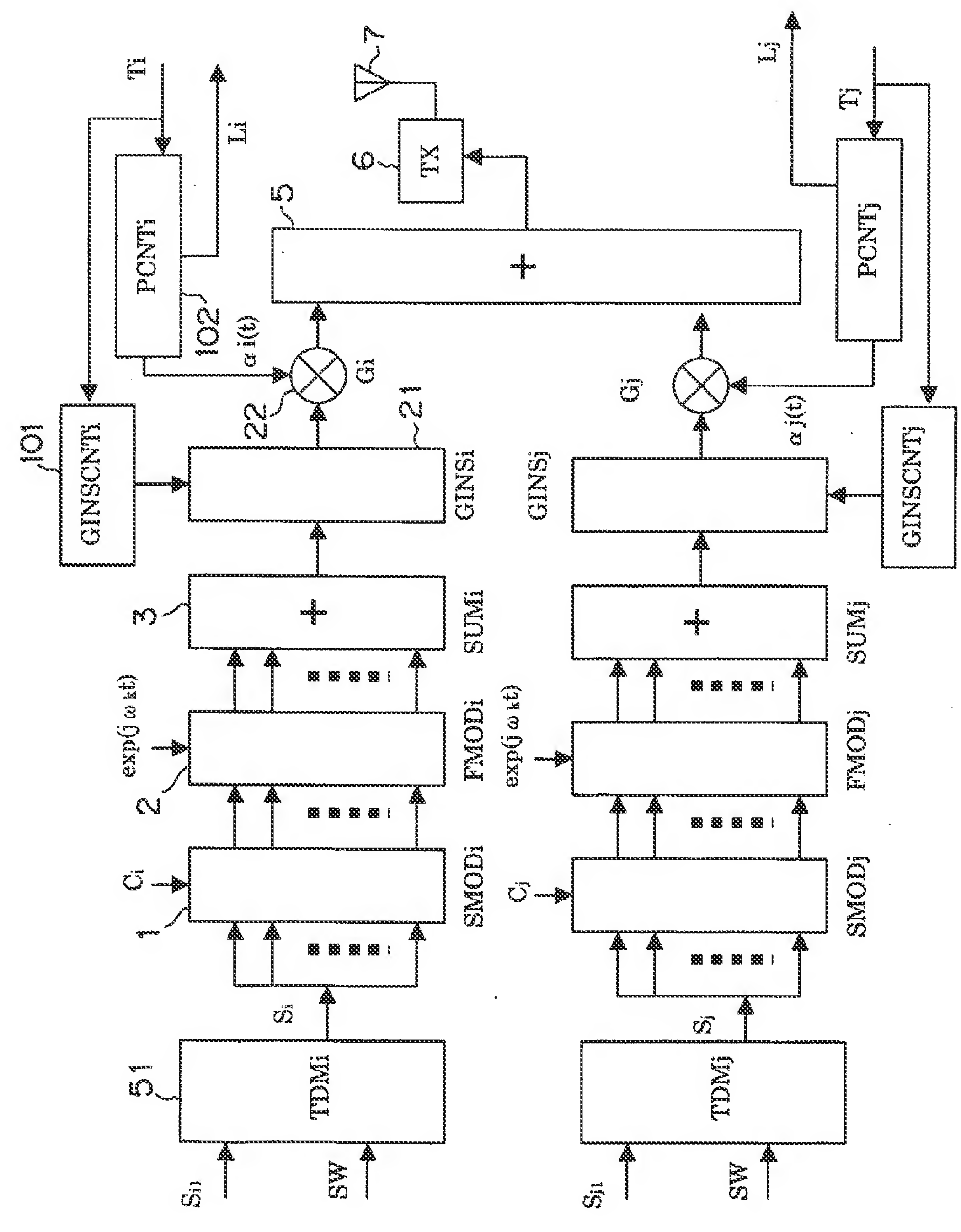

Ti
Li
PCNTi
102
101
GINSCNTi
α i(t)
Gi
22
21
GINSi
7
6
TX
5
+
Lj
Tj
PCNTj
Gj
α j(t)
GINSCNTj
GINSj
3
SUMi
SUMj
exp(jωkt)
2
FMODi
FMODj
Ci
Cj
1
SMODi
SMODj
Si
Sj
51
TDMi
TDMj
Si1
Sj1
SW

図３２

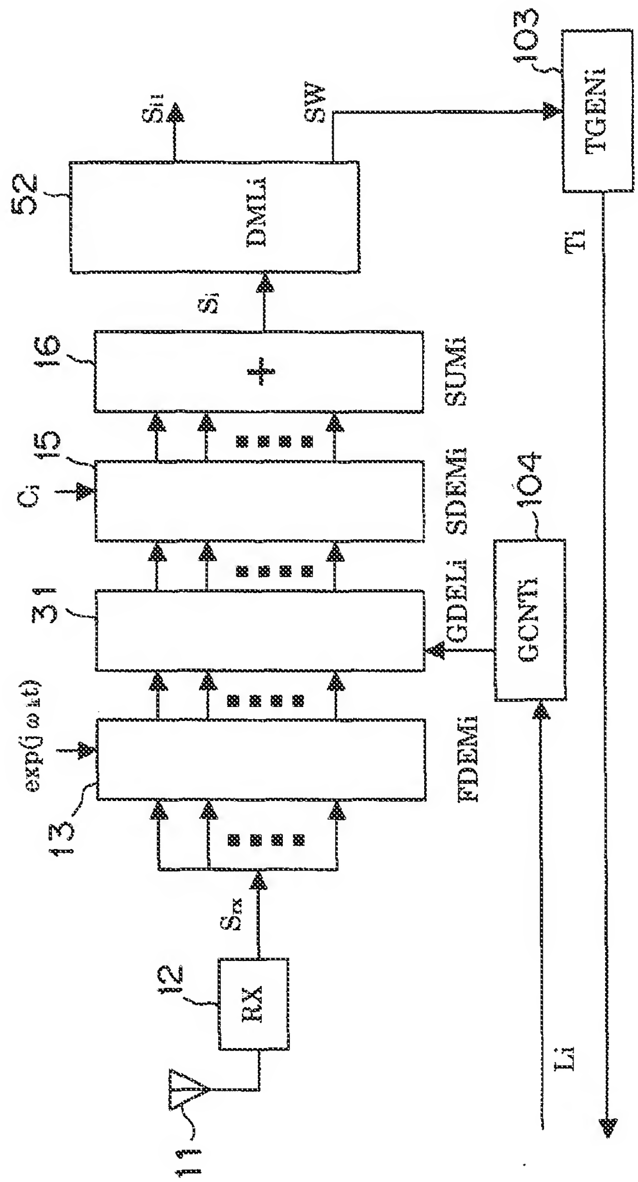

11
12
RX
S_{rx}
13
exp(jωit)
FDEMi
31
GDELi
104
GCNTi
Li
15
C_i
SDEMi
16
+
SUMi
S_i
52
DMLi
Sii
SW
103
TGENi
Ti

図33

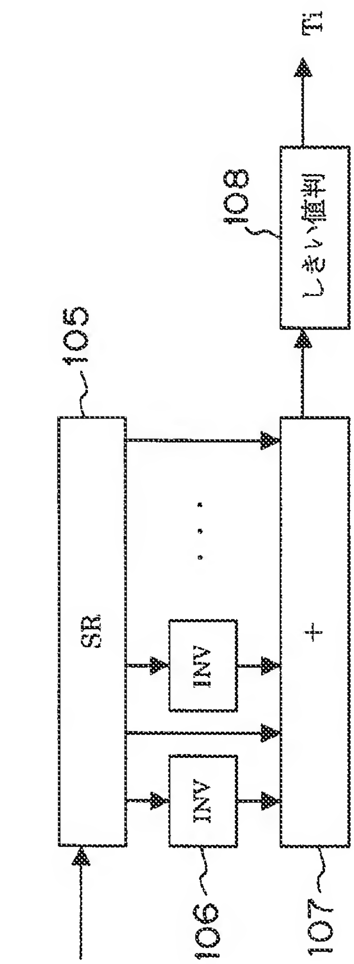


図34

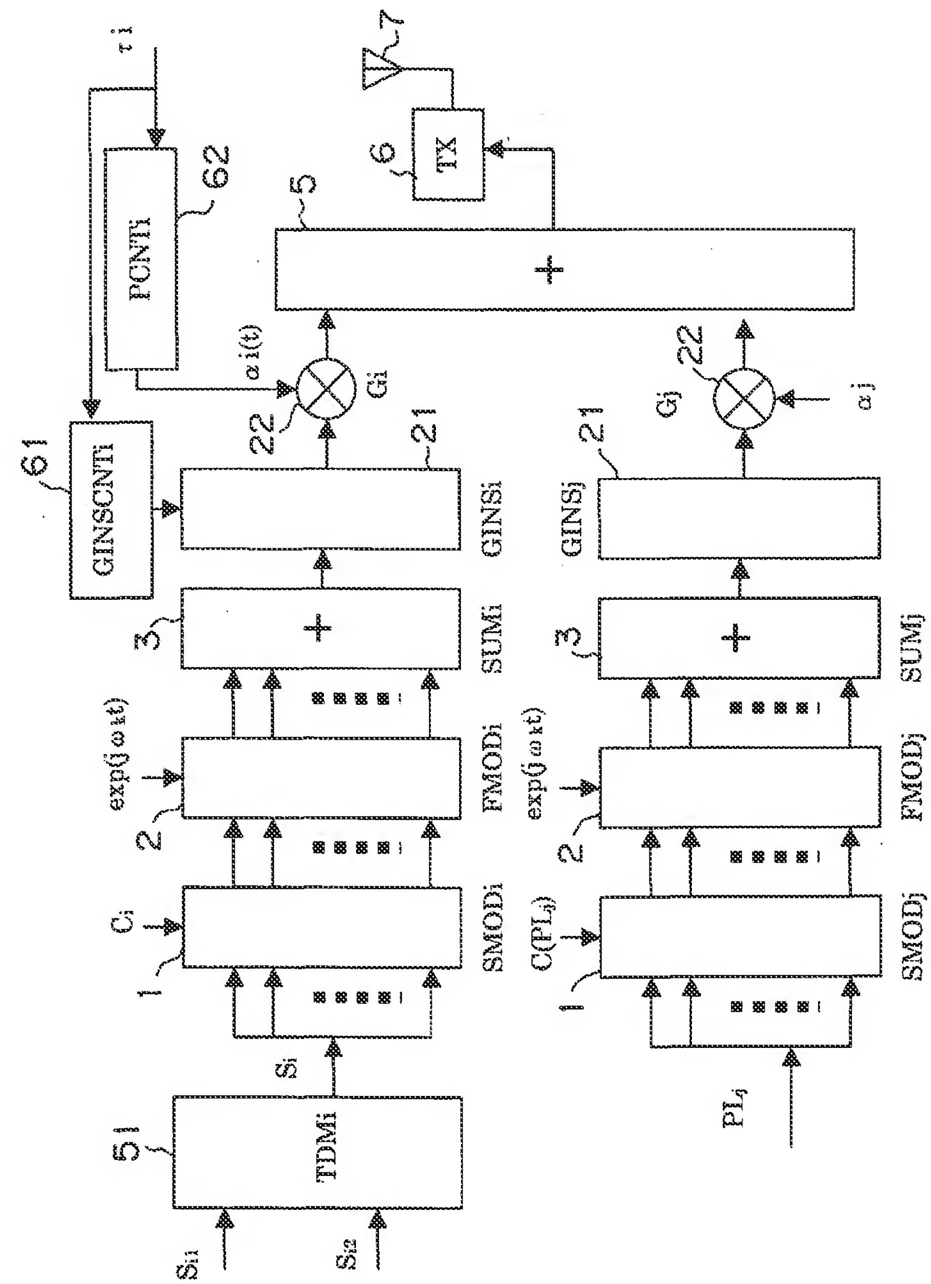

TDMi
SMODi
FMODi
SUMi
GINSi
GINSCNTi
PCNTi
TX
SMODj
FMODj
SUMj
GINSj
Gi
Gj
+

図35

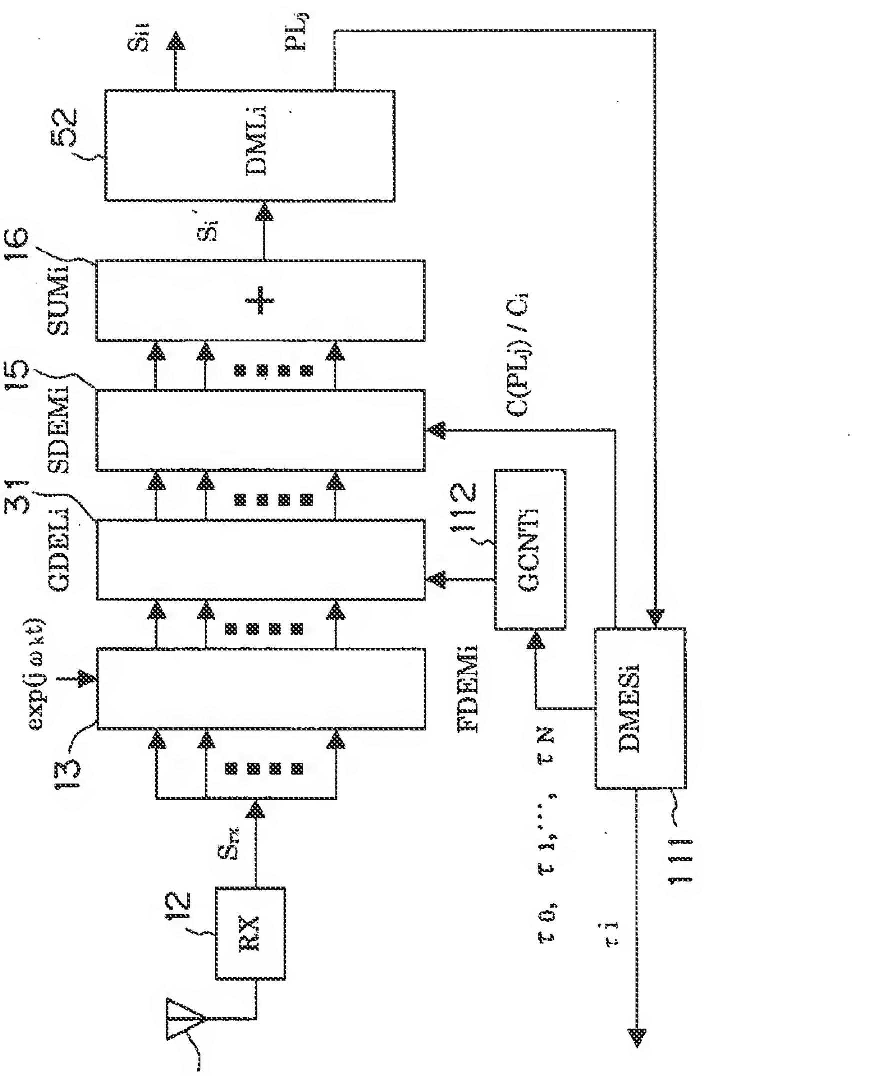

11
12
RX
Srx
13
exp(jωkt)
FDEMi
31
GDELi
15
SDEMi
16
SUMi
+
52
DMLi
Si'
Si
PLj
112
GCNTi
C(PLj) / Ci
DMESi
111
τ0, τ1,…, τN
τi

図36

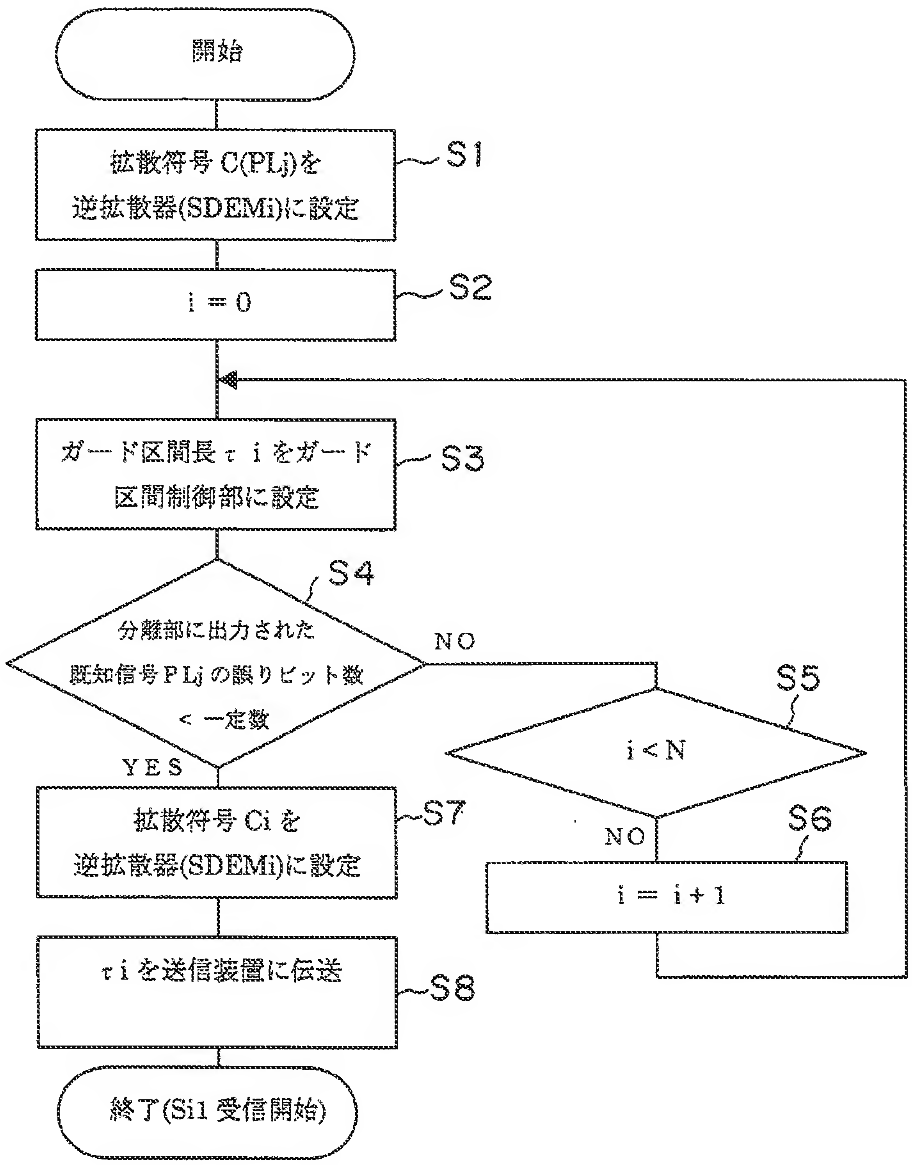

開始
拡散符号 C(PLj)を
逆拡散器(SDEMi)に設定
S1
i = 0
S2
ガード区間長τ i をガード
区間制御部に設定
S3
S4
分離部に出力された
既知信号 P Lj の誤りビット数
< 一定数
NO
YES
S5
i < N
NO
S6
i = i + 1
拡散符号 Ci を
逆拡散器(SDEMi)に設定
S7
τ i を送信装置に伝送
S8
終了(Si1 受信開始)

図37

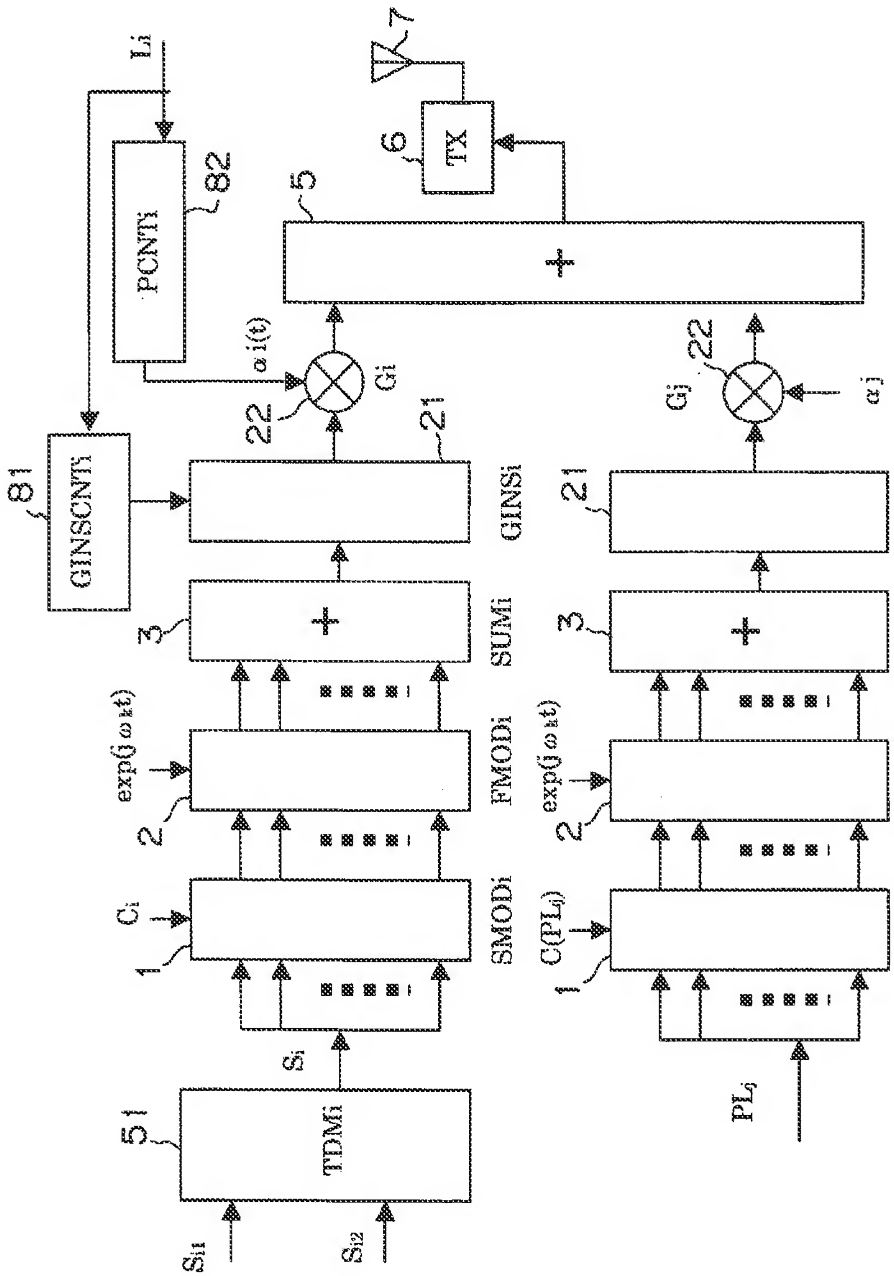
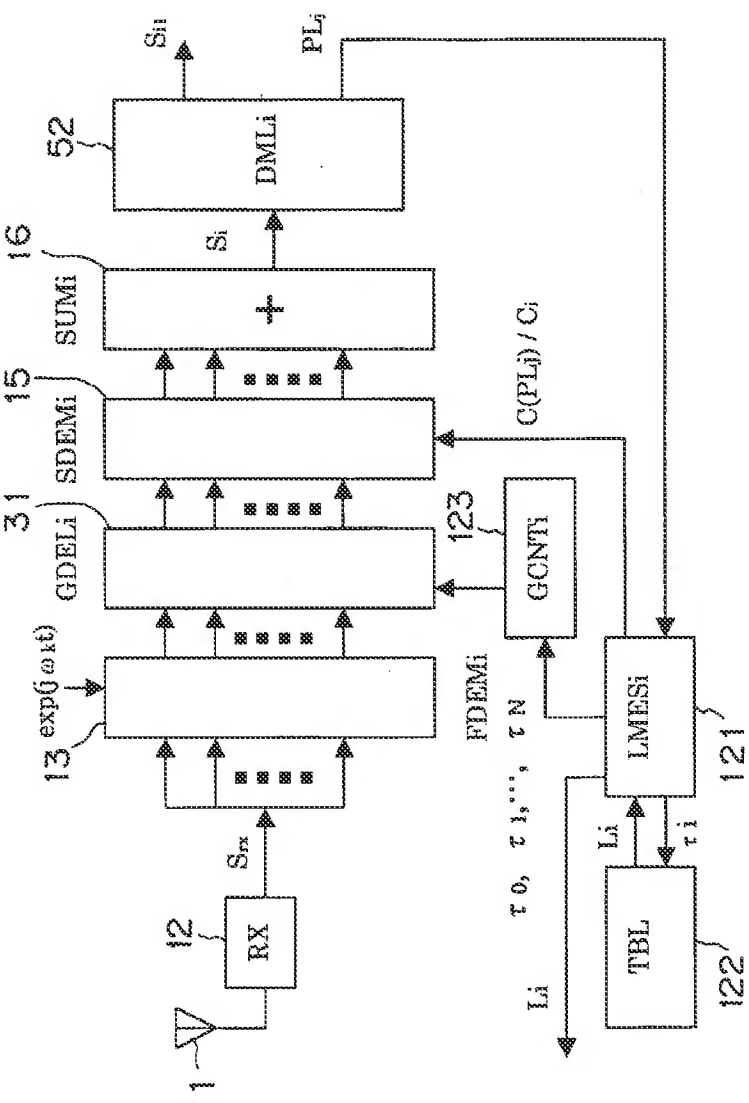

TDMi
SMODi
FMODi
SUMi
GINSi
GINSCNTi
PCNTi
TX
SMODj
FMODj
SUMj
GINSj
Gi
Gj
Li
PLj
C(PLj)
exp(jωkt)
αi(t)
αj

図38



11
12
RX
S_{rx}
13
$\exp(j\omega_i t)$
FDEMi
31
GDELi
123
GCNTi
15
SDEMi
$C(PL_j) / C_i$
16
SUMi
+
S_i
52
DMLi
S_{i1}
PL_j
121
LMESi
$\tau 0, \tau 1, \cdots, \tau N$
L_i
τi
122
TBL

図39

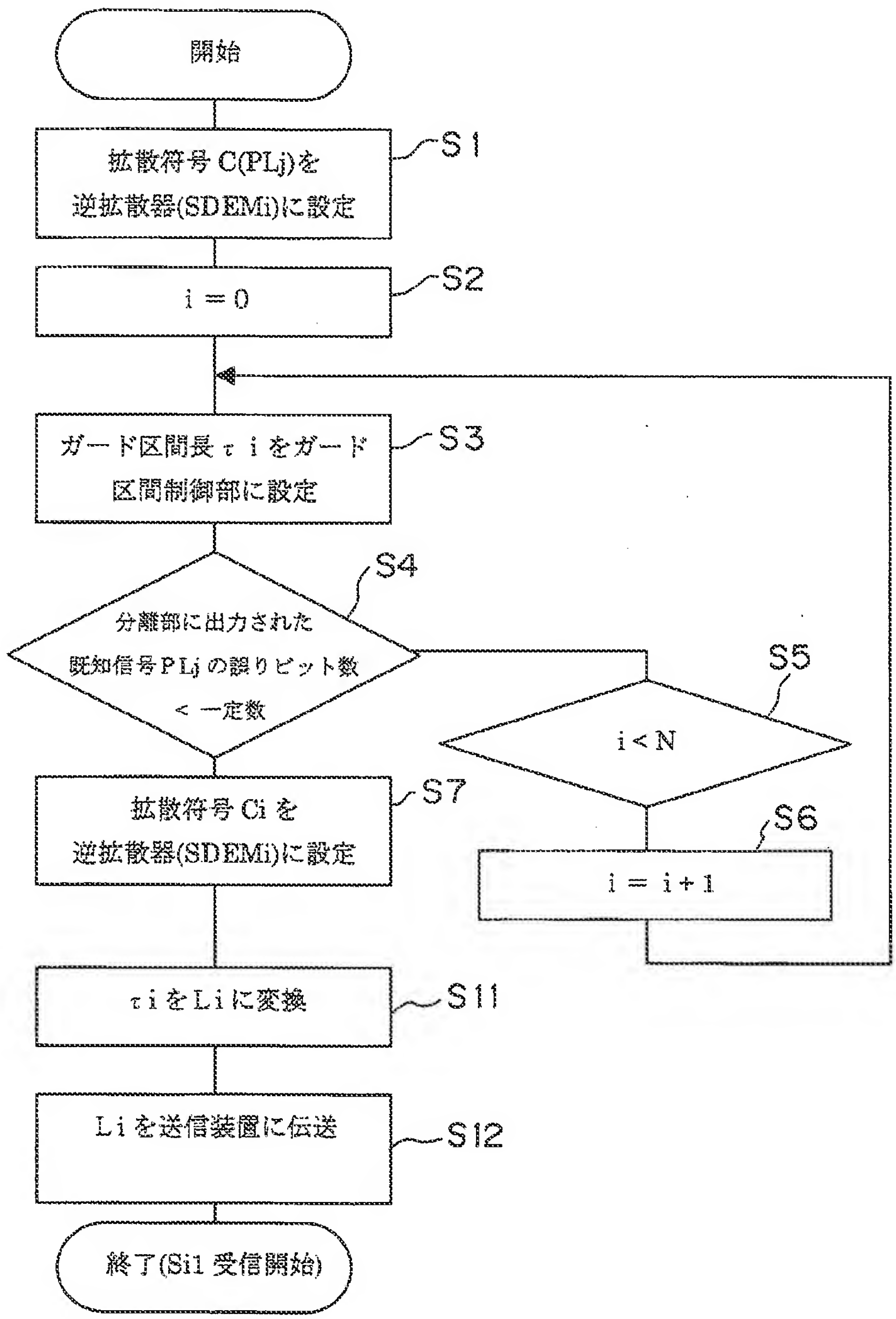

開始
拡散符号 C(PLj)を
逆拡散器(SDEMi)に設定
S1
i = 0
S2
ガード区間長τiをガード
区間制御部に設定
S3
分離部に出力された
既知信号PLjの誤りビット数
< 一定数
S4
i < N
S5
i = i+1
S6
拡散符号 Ci を
逆拡散器(SDEMi)に設定
S7
τiをLiに変換
S11
Liを送信装置に伝送
S12
終了(Si1 受信開始)

図40

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP01/10357

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl[7] H04J11/00

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl[7] H04J11/00

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
Jitsuyo Shinan Koho 1926-2000
Kokai Jitsuyo Shinan Koho 1971-2000

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	EP 1035693 A (Matsushita Electric Industrial Co., Ltd.), 13 September 2000 (13.09.2000), page 43, lines 7 to 40; Fig.146 & JP 7-99522 A, page 36, left column, line 35 to right column, line 32	1,3-5,11,13, 16,18-20
A		7-9
X	JP 2000-165342 A (Matsushita Electric Ind. Co., Ltd.), 16 June 2000 (16.06.2000), page 3, left column, line 42 to right column, line 17	1,2,6,10-12, 14-17
A	(Family: none)	7-9
X	EP 1014639 A (Matsushita Electric Industrial Co., Ltd.), 28 June 2000 (28.06.2000), page 4, lines 13 to 18 & JP 2000-244441 A, page 6, right column, lines 22 to 34 & CN 1260649 A & KR 2000052538 A	1,6,10,11, 14-16
A		7-9
A	JP 2001-111519 A (Matsushita Electric Ind. Co., Ltd.), 20 February 2001 (20.04.2001), page 3, right column, lines 9 to 17	1-20

☒ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
29 January, 2002 (29.01.02)	05 February, 2002 (05.02.02)
Name and mailing address of the ISA/ Japanese Patent Office	Authorized officer
Facsimile No.	Telephone No.

C (Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
	(Family: none)	
A	JP 11-196062 A (Advanced Digital Television Broadcasting), 21 July 1999 (21.07.1991), page 2, right column, lines 37 to 42 (Family: none)	1-20

国際調査報告

国際出願番号　ＰＣＴ／ＪＰ０１／１０３５７

Ａ．　発明の属する分野の分類（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ^{7}　Ｈ０４Ｊ１１／００

Ｂ．　調査を行った分野
調査を行った最小限資料（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ^{7}　Ｈ０４Ｊ１１／００

最小限資料以外の資料で調査を行った分野に含まれるもの
日本国実用新案公報　　　　１９２６－２０００
日本国公開実用新案公報　　１９７１－２０００

国際調査で使用した電子データベース（データベースの名称、調査に使用した用語）

Ｃ．　関連すると認められる文献

引用文献のカテゴリー＊	引用文献名　及び一部の箇所が関連するときは、その関連する箇所の表示	関連する請求の範囲の番号
Ｘ Ａ	ＥＰ　１０３５６９３　Ａ（Ｍａｔｓｕｓｈｉｔａ　Ｅｌｅｃｔｒｉｃ　Ｉｎｄｕｓｔｒｉａｌ　Ｃｏ．，　Ｌｔｄ．），２０００．０９．１３，第４３頁第７行目－第４０行目，ＦＩＧ．１４６ ＆　ＪＰ　７－９９５２２　Ａ，第３６頁左欄第３５行目－右欄第３２行目	1, 3-5, 11, 13, 16, 18-20 7-9

☒　Ｃ欄の続きにも文献が列挙されている。　　☐　パテントファミリーに関する別紙を参照。

＊　引用文献のカテゴリー
「Ａ」特に関連のある文献ではなく、一般的技術水準を示すもの
「Ｅ」国際出願日前の出願または特許であるが、国際出願日以後に公表されたもの
「Ｌ」優先権主張に疑義を提起する文献又は他の文献の発行日若しくは他の特別な理由を確立するために引用する文献（理由を付す）
「Ｏ」口頭による開示、使用、展示等に言及する文献
「Ｐ」国際出願日前で、かつ優先権の主張の基礎となる出願の日の後に公表された文献
「Ｔ」国際出願日又は優先日後に公表された文献であって出願と矛盾するものではなく、発明の原理又は理論の理解のために引用するもの
「Ｘ」特に関連のある文献であって、当該文献のみで発明の新規性又は進歩性がないと考えられるもの
「Ｙ」特に関連のある文献であって、当該文献と他の１以上の文献との、当業者にとって自明である組合せによって進歩性がないと考えられるもの
「＆」同一パテントファミリー文献

国際調査を完了した日　２９．０１．０２	国際調査報告の発送日　05.02.02
国際調査機関の名称及びあて先 日本国特許庁（ＩＳＡ／ＪＰ） 郵便番号１００－８９１５ 東京都千代田区霞が関三丁目４番３号	特許庁審査官（権限のある職員）　５Ｋ　９６４７ 高野　洋 電話番号　０３－３５８１－１１０１　内線　３５５５

Ｃ（続き）．　関連すると認められる文献

引用文献のカテゴリー＊	引用文献名　及び一部の箇所が関連するときは、その関連する箇所の表示	関連する請求の範囲の番号
X	ＪＰ　２０００－１６５３４２　Ａ（松下電器産業株式会社），２０００．０６．１６，第３頁左欄第４２行目－右欄第１７行目（ファミリーなし）	1,2,6,10-12,14-17
A		7-9
X	ＥＰ　１０１４６３９　Ａ（Ｍａｔｓｕｓｈｉｔａ　Ｅｌｅｃｔｒｉｃ　Ｉｎｄｕｓｔｒｉａｌ　Ｃｏ．，　Ｌｔｄ．），２０００．０６．２８，第４頁第１３行目－第１８行目 ＆　ＪＰ　２０００－２４４４４１　Ａ，第６頁右欄第２２行目－第３４行目 ＆　ＣＮ　１２６０６４９　Ａ ＆　ＫＲ　２０００００５２５３８　Ａ	1,6,10,11,14-16
A		7-9
A	ＪＰ　２００１－１１１５１９　Ａ（松下電器産業株式会社），２００１．０４．２０，第３頁右欄第９行目－第１７行目（ファミリーなし）	1-20
A	ＪＰ　１１－１９６０６２　Ａ（株式会社次世代デジタルテレビジョン放送システム研究所），１９９９．０７．２１，第２頁右欄第３７行目－第４２行目（ファミリーなし）	1-20

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
31 December 2003 (31.12.2003)

PCT

(10) International Publication Number
WO 2004/002047 A1

(51) International Patent Classification[7]: H04L 1/06, 25/03, 25/02

(21) International Application Number: PCT/US2003/019464

(22) International Filing Date: 20 June 2003 (20.06.2003)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
10/179,442 24 June 2002 (24.06.2002) US

(71) Applicant: QUALCOMM, INCORPORATED [US/US]; 5775 Morehouse Drive, San Diego, CA 92121 (US).

(72) Inventors: KETCHUM, John W.; 37 Candleberry Lane, Harvard, MA 01451 (US). WALTON, Jay R.; 7 Ledgewood Drive, Westford, MA 01886 (US).

(74) Agents: WADSWORTH, Philip R. et al.; 5775 Morehouse Drive, San Diego, CA 92121 (US).

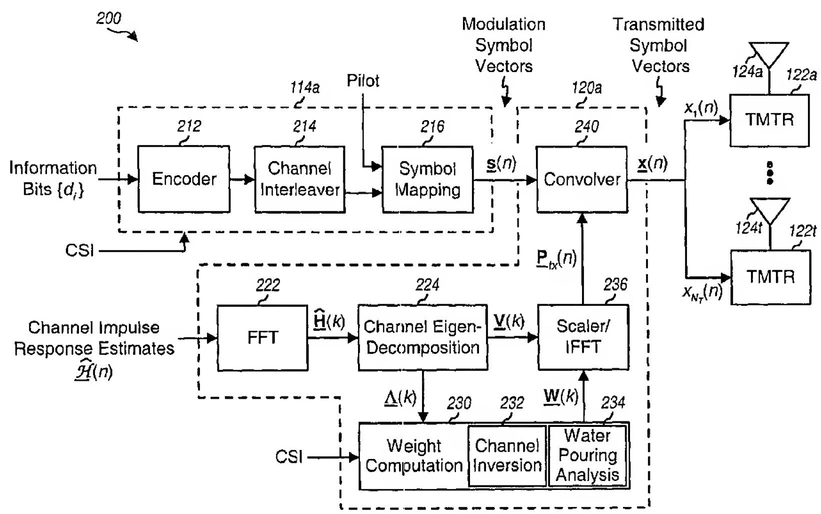
(81) Designated States (*national*): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, TJ, TM, TN, TR, TT, TZ, UA, UG, UZ, VC, VN, YU, ZA, ZM, ZW.

(84) Designated States (*regional*): ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IT, LU, MC, NL, PT, RO, SE, SI, SK, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *with international search report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) **Title:** SIGNAL PROCESSING WITH CHANNEL EIGENMODE DECOMPOSITION AND CHANNEL INVERSION FOR MIMO SYSTEMS



(57) **Abstract:** Techniques for processing a data transmission at a transmitter and receiver, which use channel eigen-decomposition, channel inversion, and (optionally) "water-pouring". At the transmitter, (1) channel eigen-decomposition is performed to determine eigenmodes of a MIMO channel and to derive a first set of steering vectors, (2) channel inversion is performed to derive weights (*e.g.*, one set for each eigenmode) used to minimize ISI distortion, and (3) water-pouring may be performed to derive scaling values indicative of the transmit powers allocated to the eigenmodes. The first set of steering vectors, weights, and scaling values are used to derive a pulse-shaping matrix, which is used to precondition modulation symbols prior to transmission. At the receiver, channel eigen-decomposition is performed to derive a second set of steering vectors, which are used to derive a pulse-shaping matrix used to condition received symbols such that orthogonal symbol streams are recovered.

WO 2004/002047 A1

SIGNAL PROCESSING WITH CHANNEL EIGENMODE DECOMPOSITION AND CHANNEL INVERSION FOR MIMO SYSTEMS

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for performing signal processing with channel eigenmode decomposition and channel inversion for multiple-input multiple-output (MIMO) communication systems.

Background

[1002] A multiple-input multiple-output (MIMO) communication system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1003] The spatial subchannels of a wideband MIMO system may encounter different channel conditions due to various factors such as fading and multipath. Each spatial subchannel may thus experience frequency selective fading, which is characterized by different channel gains at different frequencies (i.e., different frequency bins or subbands) of the overall system bandwidth. With frequency selective fading, each spatial subchannel may achieve different signal-to-noise-and-interference ratios (SNRs) for different frequency bins. Consequently, the number of information bits per modulation symbol (or data rate) that may be transmitted at different frequency bins of each spatial subchannel for a particular level of performance (e.g., 1% packet error rate) may be different from bin to bin. Moreover, because the channel conditions

typically vary with time, the supported data rates for the bins of the spatial subchannels also vary with time.

[1004] To combat frequency selective fading in a wideband channel, orthogonal frequency division multiplexing (OFDM) may be used to effectively partition the system bandwidth into a number of (N_F) subbands (which may also be referred to as frequency bins or subchannels). With OFDM, each frequency subchannel is associated with a respective subcarrier upon which data may be modulated. For a MIMO system that utilizes OFDM (i.e., a MIMO-OFDM system), each frequency subchannel of each spatial subchannel may be viewed as an independent transmission channel.

[1005] A key challenge in a coded communication system is the selection of the appropriate data rates and coding and modulation schemes to be used for a data transmission based on channel conditions. The goal of this selection process is to maximize throughput while meeting quality objectives, which may be quantified by a particular packet error rate (PER), certain latency criteria, and so on.

[1006] One straightforward technique for selecting data rates and coding and modulation schemes is to "bit load" each transmission channel in the MIMO-OFDM system according to its transmission capability, which may be quantified by the channel's short-term average SNR. However, this technique has several major drawbacks. First, coding and modulating individually for each transmission channel can significantly increase the complexity of the processing at both the transmitter and receiver. Second, coding individually for each transmission channel may greatly increase coding and decoding delay. And third, a high feedback rate would be needed to send channel state information (CSI) indicative of the channel conditions (e.g., the gain, phase, and SNR) of each transmission channel.

[1007] For a MIMO system, transmit power is another parameter that may be manipulated to maximize throughput. In general, the overall throughput of the MIMO system may be increased my allocating more transmit power to transmission channels with greater transmission capabilities. However, allocating different amounts of transmit power to different frequency bins of a given spatial subchannel tends to exaggerate the frequency selective nature of the spatial subchannel. It is well known that frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. The ISI distortion degrades performance by impacting

the ability to correctly detect the received symbols. To mitigate the deleterious effects of ISI, equalization of the received symbols would need to be performed at the receiver. Thus, a major drawback in frequency-domain power allocation is the additional complexity at the receiver to combat the resultant additional ISI distortion.

[1008] There is therefore a need in the art for techniques to achieve high overall throughput in a MIMO system without having to individually code each transmission channel and which mitigate the deleterious effects of ISI.

SUMMARY

[1009] Techniques are provided herein for processing a data transmission at a transmitter and a receiver of a MIMO system such that high performance (e.g., high overall throughput) is achieved. In an aspect, a time-domain implementation is provided which uses frequency-domain channel eigen-decomposition, channel inversion, and (optionally) "water-pouring" results to derive pulse-shaping and beam-steering solutions for the transmitter and receiver.

[1010] Channel eigen-decomposition is performed at the transmitter to determine the eigenmodes (i.e., the spatial subchannels) of a MIMO channel and to obtain a first set of steering vectors, which are used to precondition modulation symbols prior to transmission over the MIMO channel. Channel eigen-decomposition may be performed based on an estimated channel response matrix, which is an estimate of the (time-domain or frequency-domain) channel response of the MIMO channel. Channel eigen-decomposition is also performed at the receiver to obtain a second set of steering vectors, which are used to condition received symbols such that orthogonal symbol streams are recovered at the receiver.

[1011] Channel inversion is performed at the transmitter to derive weights, which are used to minimize or reduce the amount of ISI distortion at the receiver. In particular, the channel inversion may be performed for each eigenmode used for data transmission. One set of weights may be derived for each eigenmode based on the estimated channel response matrix for the MIMO channel and is used to invert the frequency response of the eigenmode.

[1012] Water-pouring analysis may (optionally) be used to more optimally allocate the total available transmit power to the eigenmodes of the MIMO channel. In particular, eigenmodes with greater transmission capabilities may be allocated more

transmit power, and eigenmodes with transmission capabilities below a particular threshold may be omitted from use (e.g., by allocating these bad eigenmodes with zero transmit power). The transmit power allocated to each eigenmode then determines the data rate and possibly the coding and modulation scheme to be used for the eigenmode.

[1013] At the transmitter, data is initially processed (e.g., coded and modulated) in accordance with a particular processing scheme to provide a number of modulation symbol streams (e.g., one modulation symbol stream for each eigenmode). An estimated channel response matrix for the MIMO channel is obtained (e.g., from the receiver) and decomposed (e.g., in the frequency domain, using channel eigen-decomposition) to obtain a set of matrices of right eigen-vectors and a set of matrices of singular values. A number of sets of weights are then derived based on the matrices of singular values, with each set of weights being used to invert the frequency response of a respective eigenmode used for data transmission. Water-pouring analysis may also be performed based on the matrices of singular values to obtain a set of scaling values indicative of the transmit powers allocated to the eigenmodes. A pulse-shaping matrix for the transmitter is then derived based on the matrices of right eigen-vectors, the weights, and the scaling values (if available). The pulse-shaping matrix comprises steering vectors, which are used to precondition the streams of modulation symbols to obtain streams of preconditioned symbols to be transmitted over the MIMO channel.

[1014] At the receiver, the estimated channel response matrix is also obtained (e.g., based on pilot symbols sent from the transmitter) and decomposed to obtain a set of matrices of left eigen-vectors. A pulse-shaping matrix for the receiver is then derived based on the matrices of left eigen-vectors and used to condition a number of received symbol streams to obtain a number of recovered symbol streams. The recovered symbols are further processed (e.g., demodulated and decoded) to recover the transmitted data.

[1015] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, digital signal processors, transmitter and receiver units, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1016] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1017] FIG. 1 is a block diagram of an embodiment of a transmitter system and a receiver system in a MIMO system;

[1018] FIG. 2 is a block diagram of a transmitter unit within the transmitter system;

[1019] FIGS. 3A and 3B are diagrams that graphically illustrate the derivation of the weights used to invert the frequency response of each eigenmode of a MIMO channel;

[1020] FIG. 4 is a flow diagram of a process for allocating the total available transmit power to the eigenmodes of the MIMO channel;

[1021] FIGS. 5A and 5B are diagrams that graphically illustrate the allocation of the total transmit power to three eigenmodes in an example MIMO system;

[1022] FIG. 6 is a flow diagram of an embodiment of the signal processing at the transmitter unit;

[1023] FIG. 7 is a block diagram of a receiver unit within the receiver system; and

[1024] FIG. 8 is a flow diagram of an embodiment of the signal processing at the receiver unit.

DETAILED DESCRIPTION

[1025] The techniques described herein for processing a data transmission at a transmitter and receiver may be used for various wireless communication systems. For clarity, various aspects and embodiments of the invention are described specifically for a multiple-input multiple-output (MIMO) communication system.

[1026] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial subchannel of the MIMO channel. The number of spatial subchannels is determined by the number of eigenmodes for the MIMO channel, which in turn is dependent on a

channel response matrix that describes the response between the N_T transmit and N_R receive antennas.

[1027] **FIG. 1** is a block diagram of an embodiment of a transmitter system 110 and a receiver system 150, which are capable of implementing various signal processing techniques described herein.

[1028] At transmitter system 110, traffic data is provided from a data source 112 to a transmit (TX) data processor 114, which formats, codes, and interleaves the traffic data based on one or more coding schemes to provide coded data. The coded traffic data may then be multiplexed with pilot data using, for example, time division multiplex (TDM) or code division multiplex (CDM), in all or a subset of the data streams to be transmitted. The pilot data is typically a known data pattern processed in a known manner, if at all. The multiplexed pilot and coded traffic data is interleaved and then modulated (i.e., symbol mapped) based on one or more modulation schemes to provide modulation symbols. In an embodiment, TX data processor 114 provides one modulation symbol stream for each spatial subchannel used for data transmission. The data rate, coding, interleaving, and modulation for each modulation symbol stream may be determined by controls provided by a controller 130.

[1029] The modulation symbols are then provided to a TX MIMO processor 120 and further processed. In a specific embodiment, the processing by TX MIMO processor 120 includes (1) determining an estimated channel frequency response matrix for the MIMO channel, (2) decomposing this matrix to determine the eigenmodes of the MIMO channel and to derive a set of "steering" vectors for the transmitter, one vector for the modulation symbol stream to be transmitted on each spatial subchannel, (3) deriving a transmit spatio-temporal pulse-shaping matrix based on the steering vectors and a weighting matrix indicative of the transmit powers assigned to the frequency bins of the eigenmodes, and (4) preconditioning the modulation symbols with the pulse-shaping matrix to provide preconditioned modulation symbols. The processing by TX MIMO processor 120 is described in further detail below. Up to N_T streams of preconditioned symbols are then provided to transmitters (TMTR) 122a through 122t.

[1030] Each transmitter 122 converts a respective preconditioned symbol stream into one or more analog signals and further conditions (e.g., amplifies, filters, and frequency upconverts) the analog signals to generate a modulated signal suitable for

transmission over the MIMO channel. The modulated signal from each transmitter 122 is then transmitted via a respective antenna 124 to the receiver system.

[1031] At receiver system 150, the transmitted modulated signals are received by N_R antennas 152a through 152r, and the received signal from each antenna 152 is provided to a respective receiver (RCVR) 154. Each receiver 154 conditions (e.g., filters, amplifies, and frequency downconverts) the received signal, digitizes the conditioned signal to provide a stream of samples, and further processes the sample stream to provide a stream of received symbols. An RX MIMO processor 160 then receives and processes the N_R received symbol streams to provide N_T streams of recovered symbols, which are estimates of the modulation symbols transmitted from the transmitter system. In an embodiment, the processing by RX MIMO processor 160 may include (1) determining the estimated channel frequency response matrix for the MIMO channel, (2) decomposing this matrix to derive a set of steering vectors for the receiver, (3) deriving a receive spatio-temporal pulse-shaping matrix based on the steering vectors, and (4) conditioning the received symbols with the pulse-shaping matrix to provide the recovered symbols. The processing by RX MIMO processor 160 is described in further detail below.

[1032] A receive (RX) data processor 162 then demodulates, deinterleaves, and decodes the recovered symbols to provide decoded data, which is an estimate of the transmitted traffic data. The processing by RX MIMO processor 160 and RX data processor 162 is complementary to that performed by TX MIMO processor 120 and TX data processor 114, respectively, at transmitter system 110.

[1033] RX MIMO processor 160 may further derive channel impulse responses for the MIMO channel, received noise power and/or signal-to-noise-and-interference ratios (SNRs) for the spatial subchannels, and so on. RX MIMO processor 160 would then provide these quantities to a controller 170. RX data processor 162 may also provide the status of each received packet or frame, one or more other performance metrics indicative of the decoded results, and possibly other information. Controller 170 then derives channel state information (CSI), which may comprise all or some of the information received from RX MIMO processor 160 and RX data processor 162. The CSI is processed by a TX data processor 178, modulated by a modulator 180, conditioned by transmitters 154a through 154r, and sent back to transmitter system 110.

[1034] At transmitter system 110, the modulated signals from receiver system 150 are received by antennas 124, conditioned by receivers 122, and demodulated by a demodulator 140 to recover the CSI transmitted by the receiver system. The CSI is then provided to controller 130 and used to generate various controls for TX data processor 114 and TX MIMO processor 120.

[1035] Controllers 130 and 170 direct the operation at the transmitter and receiver systems, respectively. Memories 132 and 172 provide storage for program codes and data used by controllers 130 and 170, respectively.

[1036] Techniques are provided herein for achieving high performance (e.g., high overall system throughput) via a time-domain implementation that uses frequency-domain channel eigen-decomposition, channel inversion, and (optionally) water-pouring results to derive time-domain pulse-shaping and beam-steering solutions for the transmitter and receiver.

[1037] Channel eigen-decomposition is performed at the transmitter to determine the eigenmodes of the MIMO channel and to derive a first set of steering vectors, which are used to precondition the modulation symbols. Channel eigen-decomposition is also performed at the receiver to derive a second set of steering vectors, which are used to condition the received symbols such that orthogonal symbol streams are recovered at the receiver. The preconditioning at the transmitter and the conditioning at the receiver orthogonalize the symbol streams transmitted over the MIMO channel.

[1038] Channel inversion is performed at the transmitter to flatten the frequency response of each eigenmode (or spatial subchannel) used for data transmission. As noted above, frequency selective fading causes intersymbol interference (ISI), which can degrade performance by impacting the ability to correctly detect the received symbols at the receiver. Conventionally, the frequency selective fading may be compensated for at the receiver by performing equalization on the received symbol streams. However, equalization increases the complexity of the receiver processing. With the inventive techniques, the channel inversion is performed at the transmitter to account for the frequency selective fading and to mitigate the need for equalization at the receiver.

[1039] Water-pouring (or water-filling) analysis is used to more optimally allocate the total available transmit power in the MIMO system to the eigenmodes such that high performance is achieved. The transmit power allocated to each eigenmode may then

determine the data rate and the coding and modulation scheme to be used for the eigenmode.

[1040] These various processing techniques are described in further detail below.

[1041] The techniques described herein provide several potential advantages. First, with time-domain eigenmode decomposition, the maximum number of eigenmodes with different SNRs is given by $\min(N_T, N_R)$. If one independent data stream is transmitted on each eigenmode and each data stream is independently processed, then the maximum number of different coding/modulation schemes is also given by $\min(N_T, N_R)$. It is also possible to make the received SNRs for the data streams essentially the same, thereby further simplifying the coding/modulation. The techniques described herein can thus greatly simplify the coding/modulation for a data transmission by avoiding the per-bin bit allocation required to approach channel capacity in MIMO-OFDM systems that utilize frequency-domain water-pouring.

[1042] Second, the channel inversion at the transmitter results in recovered symbol streams at the receiver that do not require equalization. This then reduces the complexity of the receiver processing. In contrast, other wide-band time-domain techniques typically require complicated space-time equalization to recover the symbol streams.

[1043] Third, the time-domain signaling techniques described herein can more easily integrate the channel/pilot structures of various CDMA standards, which are also based on time-domain signaling. Implementation of the channel/pilot structures may be more complicated in OFDM-based systems that perform frequency-domain signaling.

[1044] **FIG. 2** is a block diagram of an embodiment of a transmitter unit 200, which is capable of implementing various processing techniques described herein. Transmitter unit 200 is an embodiment of the transmitter portion of transmitter system 110 in FIG. 1. Transmitter unit 200 includes (1) a TX data processor 114a that receives and processes traffic and pilot data to provide N_T modulation symbol streams and (2) a TX MIMO processor 120a that preconditions the modulation symbol streams to provide N_T preconditioned symbol streams. TX data processor 114a and TX MIMO processor 120a are one embodiment of TX data processor 114 and TX MIMO processor 120, respectively, in FIG. 1.

[1045] In the specific embodiment shown in FIG. 2, TX data processor 114a includes an encoder 212, a channel interleaver 214, and a symbol mapping element 216. Encoder 212 receives and codes the traffic data (i.e., the information bits, d_i) in accordance with one or more coding schemes to provide coded bits. The coding increases the reliability of the data transmission. In an embodiment, a separate coding scheme may be used for the information bits for each eigenmode (or spatial subchannel) selected for use for data transmission. In alternative embodiments, a separate coding scheme may be used for each subset of spatial subchannels, or a common coding scheme may be used for all spatial subchannels. The coding scheme(s) to be used are determined by controls from controller 130 and may be selected based on the CSI received from the receiver system. Each selected coding scheme may include any combination of cyclic redundancy check (CRC), convolutional coding, Turbo coding, block coding, and other coding, or no coding at all.

[1046] Channel interleaver 214 interleaves the coded bits based on one or more interleaving schemes. Typically, each selected coding scheme is associated with a corresponding interleaving scheme. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average SNR of each spatial subchannel used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1047] Symbol mapping element 216 then receives and multiplexes pilot data with the interleaved data and further maps the multiplexed data in accordance with one or more modulation schemes to provide modulation symbols. A separate modulation scheme may be used for each spatial subchannel selected for use, or for each subset of spatial subchannels. Alternatively, a common modulation scheme may be used for all selected spatial subchannels.

[1048] The symbol mapping for each spatial subchannel may be achieved by grouping sets of bits to form data symbols (each of which may be a non-binary value) and mapping each data symbol to a point in a signal constellation corresponding to the modulation scheme selected for use for that spatial subchannel. The selected modulation scheme may be QPSK, M-PSK, M-QAM, or some other scheme. Each mapped signal point is a complex value and corresponds to a modulation symbol. Symbol mapping element 216 provides a vector of modulation symbols for each symbol

period, with the number of modulation symbols in each vector corresponding to the number of spatial subchannels selected for use for that symbol period. Symbol mapping element 216 thus provides up to N_T modulation symbol streams. These streams collectively form a sequence of vectors, with are also referred to as the modulation symbol vectors, $\underline{s}(n)$, with each such vector including up to N_S modulation symbols to be transmitted on up to N_S spatial subchannels for the n-th symbol period.

[1049] Within TX MIMO processor 120a, the response of the MIMO channel is estimated and used to precondition the modulation symbols prior to transmission to the receiver system. In a frequency division duplexed (FDD) system, the downlink and uplink are allocated different frequency bands, and the channel responses for the downlink and uplink may not be correlated to a sufficient degree. For the FDD system, the channel response may be estimated at the receiver and sent back to the transmitter. In a time division duplexed (TDD) system, the downlink and uplink share the same frequency band in a time division multiplexed manner, and a high degree of correlation may exist between the downlink and uplink channel responses. For the TDD system, the transmitter system may estimate the uplink channel response (e.g., based on the pilot transmitted by the receiver system on the uplink) and may then derive the downlink channel response by accounting for any differences such as those between the transmit and receive antenna array manifolds.

[1050] In an embodiment, the channel response estimates are provided to TX MIMO processor 120a as a sequence of $N_R \times N_T$ matrices, $\underline{\hat{\mathcal{H}}}(n)$, of time-domain samples. This sequence of matrices is collectively referred to as a channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$. The (i, j)-th element, $\underline{\hat{h}}_{i,j}$, of the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, for $i = (1, 2, \ldots, N_R)$ and $j = (1, 2, \ldots, N_T)$, is a sequence of samples that represents the sampled impulse response of the propagation path from the j-th transmit antenna to the i-th receive antenna.

[1051] Within TX MIMO processor 120a, a fast Fourier transformer 222 receives the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$ (e.g., from the receiver system) and derives the corresponding estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, by performing a fast Fourier transform (FFT) on the matrix $\underline{\underline{\hat{\mathcal{H}}}}$ (i.e., $\underline{\underline{\hat{\mathbf{H}}}} = \mathrm{FFT}[\underline{\underline{\hat{\mathcal{H}}}}]$). This

may be achieved by performing an N_F-point FFT on a sequence of N_F samples for each element of $\hat{\underline{\underline{\mathcal{H}}}}$ to derive a set of N_F coefficients for the corresponding element of $\hat{\underline{\underline{\mathbf{H}}}}$, where N_F corresponds to the number of frequency bins for the FFT (i.e., the length of the FFT). The $N_R \cdot N_T$ elements of $\hat{\underline{\underline{\mathbf{H}}}}$ are thus $N_R \cdot N_T$ sets of coefficients representing the frequency responses of the propagation paths between the N_T transmit antennas and N_R receive antennas. Each element of $\hat{\underline{\underline{\mathbf{H}}}}$ is the FFT of the corresponding element of $\hat{\underline{\underline{\mathcal{H}}}}$. The estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$, may also be viewed as comprising a set of N_F matrices, $\hat{\underline{\mathbf{H}}}(k)$ for $k = (0, 1, \ldots, N_F - 1)$.

Channel Eigen-Decomposition

[1052] A unit 224 then performs eigen-decomposition of the MIMO channel used for data transmission. In one embodiment for performing channel eigen-decomposition, unit 224 computes the singular value decomposition (SVD) of the estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$. In an embodiment, the singular value decomposition is performed for each matrix $\hat{\underline{\mathbf{H}}}(k)$, for $k = (0, 1, \ldots, N_F - 1)$. The singular value decomposition of matrix $\hat{\underline{\mathbf{H}}}(k)$ for frequency bin k (or frequency f_k) may be expressed as:

$$\hat{\underline{\mathbf{H}}}(k) = \underline{\mathbf{U}}(k)\underline{\Lambda}(k)\underline{\mathbf{V}}^H(k) \ , \qquad \text{Eq (1)}$$

where $\underline{\mathbf{U}}(k)$ is an $N_R \times N_R$ unitary matrix (i.e., $\underline{\mathbf{U}}^H\underline{\mathbf{U}} = \underline{\mathbf{I}}$, where $\underline{\mathbf{I}}$ is the identity matrix with ones along the diagonal and zeros everywhere else);

$\underline{\Lambda}(k)$ is an $N_R \times N_T$ diagonal matrix of singular values of $\hat{\underline{\mathbf{H}}}(k)$; and

$\underline{\mathbf{V}}(k)$ is an $N_T \times N_T$ unitary matrix.

The diagonal matrix $\underline{\Lambda}(k)$ contains non-negative real values along the diagonal (i.e., $\underline{\Lambda}(k) = \text{diag}\,(\lambda_1(k), \lambda_2(k), \ldots, \lambda_{N_T}(k))$) and zeros elsewhere. The $\lambda_i(k)$, for $i = (1, 2, \ldots, N_T)$, are referred to as the singular values of the matrix $\hat{\underline{\mathbf{H}}}(k)$. The singular value decomposition is a matrix operation known in the art and described in various references. One such reference is a book by Gilbert Strang entitled "Linear

Algebra and Its Applications," Second Edition, Academic Press, 1980, which is incorporated herein by reference.

[1053] The result of the singular value decomposition is three sets of N_F matrices, $\underline{\underline{\mathbf{U}}}$, $\underline{\underline{\Lambda}}$, and $\underline{\underline{\mathbf{V}}}^H$, where $\underline{\underline{\mathbf{U}}} = [\underline{\mathbf{U}}(0) \ \ldots \ \underline{\mathbf{U}}(k) \ \ldots \ \underline{\mathbf{U}}(N_F - 1)]$, and so on. For each value of k, $\underline{\mathbf{U}}(k)$ is the $N_R \times N_R$ unitary matrix of left eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, $\underline{\mathbf{V}}(k)$ is the $N_T \times N_T$ unitary matrix of right eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, and $\underline{\Lambda}(k)$ is the $N_R \times N_T$ diagonal matrix of singular values of $\hat{\underline{\mathbf{H}}}(k)$.

[1054] In another embodiment for performing channel eigen-decomposition, unit 224 first obtains a square matrix $\underline{\mathbf{R}}(k)$ as $\underline{\mathbf{R}}(k) = \hat{\underline{\mathbf{H}}}^H(k)\hat{\underline{\mathbf{H}}}(k)$. The eigenvalues of the square matrix $\underline{\mathbf{R}}(k)$ would then be the squares of the singular values of the matrix $\hat{\underline{\mathbf{H}}}(k)$, and the eigen-vectors of $\underline{\mathbf{R}}(k)$ would be the right eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, or $\underline{\mathbf{V}}(k)$. The decomposition of $\underline{\mathbf{R}}(k)$ to obtain the eigenvalues and eigen-vectors is known in the art and not described herein. Similarly, another square matrix $\underline{\mathbf{R}}'(k)$ may be obtained as $\underline{\mathbf{R}}'(k) = \hat{\underline{\mathbf{H}}}(k)\hat{\underline{\mathbf{H}}}^H(k)$. The eigenvalues of this square matrix $\underline{\mathbf{R}}'(k)$ would also be the squares of the singular values of $\hat{\underline{\mathbf{H}}}(k)$, and the eigen-vectors of $\underline{\mathbf{R}}'(k)$ would be the left eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, or $\underline{\mathbf{U}}(k)$.

[1055] The channel eigen-decomposition is used to decompose the MIMO channel into its eigenmodes, at frequency f_k, for each value of k where $k = (0, 1, \ldots, N_F - 1)$. The rank $r(k)$ of $\hat{\underline{\mathbf{H}}}(k)$ corresponds to the number of eigenmodes for the MIMO channel at frequency f_k, which corresponds to the number of independent channels (i.e., the number of spatial subchannels) available in frequency bin k.

[1056] As described in further detail below, the columns of $\underline{\mathbf{V}}(k)$ are the steering vectors associated with frequency f_k to be used at the transmitter for the elements of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. Correspondingly, the columns of $\underline{\mathbf{U}}(k)$ are the steering vectors associated with frequency f_k to be used at the receiver for the elements of the received symbol vectors, $\underline{\mathbf{r}}(n)$. The matrices $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$, for $k = (0, 1, \ldots, N_F - 1)$, are used to orthogonalize the symbol streams transmitted on the

eigenmodes at each frequency f_k. When these matrices are used to precondition the modulation symbol streams at the transmitter and to condition the received symbol streams at the receiver, either in the frequency domain or the time domain, the result is the overall orthogonalization of the symbol streams. This then allows for separate coding/modulation per eigenmode (as opposed to per bin), which can greatly simplify the processing at the transmitter and receiver.

[1057] The elements along the diagonal of $\underline{\Lambda}(k)$ are $\lambda_{ii}(k)$, for $i = \{1,\ 2,\ \ldots,\ r(k)\}$, where $r(k)$ is the rank of $\hat{\underline{\mathbf{H}}}(k)$. The columns of $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$, $\underline{\mathbf{u}}_i(k)$ and $\underline{\mathbf{v}}_i(k)$, respectively, are solutions to the eigen equation, which may be expressed as:

$$\hat{\underline{\mathbf{H}}}(k)\underline{\mathbf{v}}_i(k) = \lambda_{ii}\underline{\mathbf{u}}_i(k) \quad . \qquad \text{Eq (2)}$$

[1058] The three sets of matrices, $\underline{\mathbf{U}}(k)$, $\underline{\Lambda}(k)$, and $\underline{\mathbf{V}}(k)$, for $k = (0,\ 1,\ \ldots,\ N_F - 1)$, may be provided in two forms - a "sorted" form and a "random-ordered" form. In the sorted form, the diagonal elements of each matrix $\underline{\Lambda}(k)$ are sorted in decreasing order so that $\lambda_{11}(k) \geq \lambda_{22}(k) \geq \ldots \geq \lambda_{rr}(k)$, and their eigen-vectors are arranged in corresponding order in $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$. The sorted form is indicated by the subscript *s*, i.e., $\underline{\mathbf{U}}_s(k)$, $\underline{\Lambda}_s(k)$, and $\underline{\mathbf{V}}_s(k)$, for $k = (0,\ 1,\ \ldots,\ N_F - 1)$.

[1059] In the random-ordered form, the ordering of the singular values and eigen-vectors may be random and further independent of frequency. The random form is indicated by the subscript *r*. The particular form selected for use, sorted or random-ordered, influences the selection of the eigenmodes for use for data transmission and the coding and modulation scheme to be used for each selected eigenmode.

[1060] A weight computation unit 230 receives the set of diagonal matrices, $\underline{\underline{\Lambda}}$, which contains a set of singular values (i.e., $\lambda_{11}(k),\ \lambda_{22}(k),\ \ldots,\lambda_{rr}(k)$) for each frequency bin. Weight computation unit 230 then derives a set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, where $\underline{\underline{\mathbf{W}}} = [\underline{\mathbf{W}}(0)\ \ldots\ \underline{\mathbf{W}}(k)\ \ldots\ \underline{\mathbf{W}}(N_F - 1)]$. The weighting matrices are used to scale the modulation symbol vectors, $\underline{\mathbf{s}}(n)$, in either the time or frequency domain, as described below.

[1061] Weight computation unit 230 includes a channel inversion unit 232 and (optionally) a water-pouring analysis unit 234. Channel inversion unit 232 derives a set of weights, $\underline{\mathbf{w}}_{ii}$, for each eigenmode, which is used to combat the frequency selective fading on the eigenmode. Water-pouring analysis unit 234 derives a set of scaling values, $\underline{\mathbf{b}}$, for the eigenmodes of the MIMO channel. These scaling values are indicative of the transmit powers allocated to the eigenmodes. Channel inversion and water-pouring are described in further detail below.

Channel Inversion

[1062] **FIG. 3A** is a diagram that graphically illustrates the derivation of the set of weights, $\underline{\mathbf{w}}_{ii}$, used to invert the frequency response of each eigenmode. The set of diagonal matrices, $\underline{\Lambda}(k)$ for $k = (0, 1, \ldots, N_F - 1)$, is shown arranged in order along an axis 310 that represents the frequency dimension. The singular values, $\lambda_{ii}(k)$ for $i = (1, 2, \ldots, N_S)$, of each matrix $\underline{\Lambda}(k)$ are located along the diagonal of the matrix. Axis 312 may thus be viewed as representing the spatial dimension. Each eigenmode of the MIMO channel is associated with a set of elements, $\{\lambda_{ii}(k)\}$ for $k = (0, 1, \ldots, N_F - 1)$, that is indicative of the frequency response of that eigenmode. The set of elements $\{\lambda_{ii}(k)\}$ for each eigenmode is shown by the shaded boxes along a dashed line 314. For each eigenmode that experiences frequency selective fading, the elements $\{\lambda_{ii}(k)\}$ for the eigenmode may be different for different values of k.

[1063] Since frequency selective fading causes ISI, the deleterious effects of ISI may be mitigated by "inverting" each eigenmode such that it appears flat in frequency at the receiver. The channel inversion may be achieved by deriving a set of weights, $\{w_{ii}(k)\}$ for $k = (0, 1, \ldots, N_F - 1)$, for each eigenmode such that the product of the weights and the corresponding eigenvalues (i.e., the squares of the diagonal elements) are approximately constant for all values of k, which may be expressed as $w_{ii}(k) \cdot \lambda_{ii}^2(k) = a_i$, for $k = (0, 1, \ldots, N_F - 1)$.

[1064] For eigenmode i, the set of weights for the N_F frequency bins, $\underline{\mathbf{w}}_{ii} = [w_{ii}(0) \ \ldots \ w_{ii}(k) \ \ldots \ w_{ii}(N_F - 1)]^T$, used to invert the channel may be derived as:

$$w_{ii}(k) = \frac{a_i}{\lambda_{ii}^2(k)} \quad , \qquad \text{for } k = (0,\ 1,\ \ldots\ ,N_F - 1)\ , \qquad \text{Eq (3)}$$

where a_i is a normalization factor that may be expressed as:

$$a_i = \frac{\sum_{k=0}^{N_F-1} \lambda_{ii}^2(k)}{\sum_{k=0}^{N_F-1} \frac{1}{\lambda_{ii}^2(k)}} \quad . \qquad \text{Eq (4)}$$

As shown in equation (4), a normalization factor a_i is determined for each eigenmode based on the set of eigenvalues (i.e., the squared singular values), $\{\lambda_{ii}^2(k)\}$ for $k = (0,\ 1,\ \ldots\ ,N_F - 1)$, associated with that eigenmode. The normalization factor a_i is defined such that $\sum_{k=0}^{N_F-1} w_{ii}(k) = \sum_{k=0}^{N_F-1} \lambda_{ii}^2(k)$.

[1065] **FIG. 3B** is a diagram that graphically illustrates the relationship between the set of weights for a given eigenmode and the set of eigenvalues for that eigenmode. For eigenmode *i*, the weight $w_{ii}(k)$ for each frequency bin is inversely related to the eigenvalue $\lambda_{ii}^2(k)$ for that bin, as shown in equation (3). To flatten the spatial subchannel and minimize or reduce ISI, it is undesirable to selectively eliminate transmit power on any frequency bin. The set of N_F weights for each eigenmode is used to scale the modulation symbols, $\underline{s}(n)$, in the frequency or time domain, prior to transmission on the eigenmode.

[1066] For the sorted order form, the singular values $\lambda_{ii}(k)$, for $i = (1,\ 2,\ \ldots\ ,N_S)$, for each matrix $\underline{\Lambda}(k)$ are sorted such that the diagonal elements of $\underline{\Lambda}(k)$ with smaller indices are generally larger. Eigenmode 0 (which is often referred to as the principle eigenmode) would then be associated with the largest singular value in each of the N_F diagonal matrices, $\underline{\Lambda}(k)$, eigenmode 1 would then be associated with the second largest singular value in each of the N_F diagonal matrices, and so on. Thus, even though the channel inversion is performed over all N_F frequency bins for each eigenmode, the eigenmodes with lower indices are not likely to have too many bad bins (if any). Thus,

at least for eigenmodes with lower indices, excessive transmit power is not used for bad bins.

[1067] The channel inversion may be performed in various manners to invert the MIMO channel, and this is within the scope of the invention. In one embodiment, the channel inversion is performed for each eigenmode selected for use. In another embodiment, the channel inversion may be performed for some eigenmodes but not others. For example, the channel inversion may be performed for each eigenmode determined to induce excessive ISI. The channel inversion may also be dynamically performed for some or all eigenmodes selected for use, for example, when the MIMO channel is determined to be frequency selective (e.g., based on some defined criteria).

[1068] Channel inversion is described in further detail in U.S. Patent Application Serial No. 09/860,274, filed May 17, 2001, U.S. Patent Application Serial No. 09/881,610, filed June 14, 2001, and U.S. Patent Application Serial No. 09/892,379, filed June 26, 2001, all three entitled "Method and Apparatus for Processing Data for Transmission in a Multi-Channel Communication System Using Selective Channel Inversion," assigned to the assignee of the present application and incorporated herein by reference.

Water-Pouring

[1069] In an embodiment, water-pouring analysis is performed (if at all) across the spatial dimension such that more transmit power is allocated to eigenmodes with better transmission capabilities. The water-pouring power allocation is analogous to pouring a fixed amount of water into a vessel with an irregular bottom, where each eigenmode corresponds to a point on the bottom of the vessel, and the elevation of the bottom at any given point corresponds to the inverse of the SNR associated with that eigenmode. A low elevation thus corresponds to a high SNR and, conversely, a high elevation corresponds to a low SNR. The total available transmit power, P_{total}, is then "poured" into the vessel such that the lower points in the vessel (i.e., those with higher SNRs) are filled first, and the higher points (i.e., those with lower SNRs) are filled later. A constant P_{set} is indicative of the water surface level for the vessel after all of the total available transmit power has been poured. This constant may be estimated initially based on various system parameters. The power allocation is dependent on the total available transmit power and the depth of the vessel over the bottom surface. The

points with elevations above the water surface level are not filled (i.e., eigenmodes with SNRs below a particular value are not used for data transmission).

[1070] In an embodiment, the water-pouring is not performed across the frequency dimension because this tends to exaggerate the frequency selectivity of the eigenmodes created by the channel eigenmode decomposition described above. The water-pouring may be performed such that all eigenmodes are used for data transmission, or only a subset of the eigenmodes is used (with bad eigenmodes being discarded). It can be shown that water-pouring across the eigenmodes, when used in conjunction with the channel inversion with the singular values sorted in descending order, can provide near-optimum performance while mitigating the need for equalization at the receiver.

[1071] The water-pouring may be performed by water-pouring analysis unit 234 as follows. Initially, the total power in each eigenmode is determined as:

$$P_{i,\lambda} = \sum_{k=0}^{N_F-1} \lambda_{ii}^2(k) \quad . \qquad \text{Eq (5)}$$

[1072] The SNR for each eigenmode may then be determined as:

$$\text{SNR}_i = \frac{P_{i,\lambda}}{\sigma^2} \quad , \qquad \text{Eq (6)}$$

where σ^2 is the received noise variance, which may also be denoted as the received noise power N_0. The received noise power corresponds to the noise power on the recovered symbols at the receiver, and is a parameter that may be provided by the receiver to the transmitter as part of the reported CSI.

[1073] The transmit power, P_i, to be allocated to each eigenmode may then be determined as:

$$P_i = \max\left[\left(P_{set} - \frac{1}{\text{SNR}_i}\right),\ 0\right] \quad , \text{ and} \qquad \text{Eq (7a)}$$

$$P_{total} \geq \sum_{i=1}^{N_S} P_i \quad , \qquad \text{Eq (7b)}$$

where P_{set} is a constant that may be derived from various system parameters, and P_{total} is the total transmit power available for allocation to the eigenmodes.

[1074] As shown in equation (7a), each eigenmode of sufficient quality is allocated transmit power of $\left(P_{set} - \frac{1}{\text{SNR}_i}\right)$. Thus, eigenmodes that achieve better SNRs are allocated more transmit powers. The constant P_{set} determines the amounts of transmit power to allocate to the better eigenmodes. This then indirectly determines which eigenmodes get selected for use since the total available transmit power is limited and the power allocation is constrained by equation (7b).

[1075] Water-pouring analysis unit 234 thus receives the set of diagonal matrices, $\underline{\underline{\Lambda}}$, and the received noise power, σ^2. The matrices $\underline{\underline{\Lambda}}$ are then used in conjunction with the received noise power to derive a vector of scaling values, $\underline{\mathbf{b}} = [b_0 \ \ldots \ b_i \ \ldots \ b_{N_S}]^T$, where $b_i = P_i$ for $i = (1, \ 2, \ \ldots \ , N_S)$. The P_i are the solutions to the water-pouring equations (7a) and (7b). The scaling values in $\underline{\mathbf{b}}$ are indicative of the transmit powers allocated to the N_S eigenmodes, where zero or more eigenmodes may be allocated no transmit power.

[1076] **FIG. 4** is a flow diagram of an embodiment of a process 400 for allocating the total available transmit power to a set of eigenmodes. Process 400, which is one specific water-pouring implementation, determines the transmit powers, P_i, for $i \in I$, to be allocated to the eigenmodes in set I, given the total transmit power, P_{total}, available at the transmitter, the set of eigenmode total powers, $P_{i,\lambda}$, and the received noise power, σ^2.

[1077] Initially, the variable n used to denote the iteration number is set to one (i.e., $n = 1$) (step 412). For the first iteration, set $I(n)$ is defined to include all of the eigenmodes for the MIMO channel, or $I(n) = \{1, \ 2, \ \ldots \ , N_S\}$ (step 414). The cardinality (or length) of set $I(n)$ for the current iteration n is then determined as $L_I(n) = |I(n)|$, which is $L_I(n) = N_S$ for the first iteration (step 416).

[1078] The total effective power, $P_{eff}(n)$, to be distributed across the eigenmodes in set $I(n)$ is next determined (step 418). The total effective power is defined to be equal

to the total available transmit power, P_{total}, plus the sum of the inverse SNRs for the eigenmodes in set $I(n)$. This may be expressed as:

$$P_{eff}(n) = P_{total} + \sum_{i \in I(n)} \frac{\sigma^2}{P_{i,\lambda}} \quad . \qquad \text{Eq (8)}$$

[1079] The total available transmit power is then allocated to the eigenmodes in set $I(n)$. The index i used to iterate through the eigenmodes in set $I(n)$ is initialized to one (i.e., $i = 1$) (step 420). The amount of transmit power to allocate to eigenmode i is then determined (step 422) based on the following:

$$P_i(n) = \frac{P_{eff}(n)}{L_I(n)} - \frac{\sigma^2}{P_{i,\lambda}} \quad . \qquad \text{Eq (9)}$$

Each eigenmode in set $I(n)$ is allocated transmit power, P_i, in step 422. Steps 424 and 426 are part of a loop to allocate transmit power to each of the eigenmodes in set $I(n)$.

[1080] **FIG. 5A** graphically illustrates the total effective power, P_{eff}, for an example MIMO system with three eigenmodes. Each eigenmode has an inverse SNR equal to $\sigma^2 / \lambda_{ii}^2$, for $i = \{1, 2, 3\}$, which assumes a normalized transmit power of 1.0. The total transmit power available at the transmitter is $P_{total} = P_1 + P_2 + P_3$, and is represented by the shaded area in FIG. 5A. The total effective power is represented by the area in the shaded and unshaded regions in FIG. 5A.

[1081] For water-pouring, although the bottom of the vessel has an irregular surface, the water level at the top remains constant across the vessel. Likewise and as shown in FIG. 5A, after the total available transmit power, P_{total}, has been distributed to the eigenmodes, the final power level is constant across all eigenmodes. This final power level is determined by dividing $P_{eff}(n)$ by the number of eigenmodes in set $I(n)$, $L_I(n)$. The amount of power allocated to eigenmode i is then determined by subtracting the inverse SNR of that eigenmode, $\sigma^2 / \lambda_{ii}^2$, from the final power level, $P_{eff}(n) / L_I(n)$, as given by equation (9) and shown in FIG. 5A.

[1082] **FIG. 5B** shows a situation whereby the water-pouring power allocation results in an eigenmode receiving negative power. This occurs when the inverse SNR

of the eigenmode is above the final power level, which is expressed by the condition $(P_{eff}(n)/L_I(n)) < (\sigma^2/\lambda_{ii}^2)$.

[1083] Referring back to FIG. 4, at the end of the power allocation, a determination is made whether or not any eigenmode has been allocated negative power (i.e., $P_i < 0$) (step 428). If the answer is yes, then the process continues by removing from set $I(n)$ all eigenmodes that have been allocated negative powers (step 430). The index n is incremented by one (i.e., $n = n+1$) (step 432). The process then returns to step 416 to allocate the total available transmit power among the remaining eigenmodes in set $I(n)$. The process continues until all eigenmodes in set $I(n)$ have been allocated positive transmit powers, as determined in step 428. The eigenmodes not in set $I(n)$ are allocated zero power.

[1084] Water-pouring is also described by Robert G. Gallager, in "Information Theory and Reliable Communication," John Wiley and Sons, 1968, which is incorporated herein by reference. A specific algorithm for performing the basic water-pouring process for a MIMO-OFDM system is described in U.S. Patent Application Serial No. 09/978,337, entitled "Method and Apparatus for Determining Power Allocation in a MIMO Communication System," filed October 15, 2001. Water-pouring is also described in further detail in U.S. Patent Application Serial No. 10/056,275, entitled "Reallocation of Excess Power for Full Channel-State Information (CSI) Multiple-Input, Multiple-Output (MIMO) Systems," filed January 23, 2002. These applications are assigned to the assignee of the present application and incorporated herein by reference.

[1085] If water-pouring is performed to allocate the total available transmit power to the eigenmodes, then water-pouring analysis unit 234 provides a set of N_S scaling values, $\underline{\mathbf{b}} = \{b_0 \; ... \; b_i \; ... \; b_{N_S}\}$, for the N_S eigenmodes. Each scaling value is for a respective eigenmode and is used to scale the set of weights determined for that eigenmode.

[1086] For eigenmode i, a set of weights, $\underline{\hat{\mathbf{w}}}_{ii} = [\hat{w}_{ii}(0) \; ... \; \hat{w}_{ii}(k) \; ... \; \hat{w}_{ii}(N_F - 1)]^T$, used to invert the channel and scale the transmit power of the eigenmode may be derived as:

$$\hat{w}_{ii}(k) = \frac{a_i b_i}{\lambda_{ii}^2(k)} \ , \qquad \text{for } k = (0,\ 1,\ \ldots\ , N_F - 1) \ , \qquad \text{Eq (10)}$$

where the normalization factor, a_i, and the scaling value, b_i, are derived as described above.

[1087] Weight computation unit 230 provides the set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, which may be obtained using the weights $w_{ii}(k)$ or $\hat{w}_{ii}(k)$. Each weighting matrix, $\underline{\mathbf{W}}(k)$, is a diagonal matrix whose diagonal elements are composed of the weights derived above. In particular, if only channel inversion is performed, then each weighting matrix, $\underline{\mathbf{W}}(k)$, for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, is defined as:

$$\underline{\mathbf{W}}(k) = \text{diag}\,(w_{11}(k),\ w_{22}(k),\ \ldots\ , w_{N_S N_S}(k)) \ , \qquad \text{Eq (11a)}$$

where $w_{ii}(k)$ is derived as shown in equation (3). And if both channel inversion and water-pouring are performed, then each weighting matrix, $\underline{\mathbf{W}}(k)$, for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, is defined as:

$$\underline{\mathbf{W}}(k) = \text{diag}\,(\hat{w}_{11}(k),\ \hat{w}_{22}(k),\ \ldots\ , \hat{w}_{N_S N_S}(k)) \ , \qquad \text{Eq (11b)}$$

where $\hat{w}_{ii}(k)$ is derived as shown in equation (10).

[1088] Referring back to FIG. 2, a scaler/IFFT 236 receives (1) the set of unitary matrices, $\underline{\underline{\mathbf{V}}}$, which are the matrices of right eigen-vectors of $\underline{\underline{\hat{\mathbf{H}}}}$, and (2) the set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, for all N_F frequency bins. Scaler/IFFT 236 then derives a spatio-temporal pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, for the transmitter based on the received matrices. Initially, the square root of each weighting matrix, $\underline{\mathbf{W}}(k)$, is computed to obtain a corresponding matrix, $\sqrt{\underline{\mathbf{W}}(k)}$, whose elements are the square roots of the elements of $\underline{\mathbf{W}}(k)$. The elements of the weighting matrices, $\underline{\mathbf{W}}(k)$ for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, are related to the power of the eigenmodes. The square root then transforms the power to the equivalent signal scaling. For each frequency bin k, the product of the square-root weighting matrix, $\sqrt{\underline{\mathbf{W}}(k)}$, and the corresponding unitary

matrix, $\underline{\mathbf{V}}(k)$, is then computed to provide a product matrix, $\underline{\mathbf{V}}(k)\sqrt{\underline{\mathbf{W}}(k)}$. The set of product matrices, $\underline{\mathbf{V}}(k)\sqrt{\underline{\mathbf{W}}(k)}$ for $k = (0, 1, \ldots, N_F - 1)$, which is also denoted as $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$, defines the optimal or near-optimal spatio-spectral shaping to be applied to the modulation symbol vectors, $\underline{\mathbf{s}}(n)$.

[1089] An inverse FFT of $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$ is then computed to derive the spatio-temporal pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, for the transmitter, which may be expressed as:

$$\underline{\mathbf{P}}_{tx}(n) = \mathrm{IFFT}\,[\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}] \quad . \qquad \text{Eq (12)}$$

The pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, is an $N_T \times N_T$ matrix. Each element of $\underline{\mathbf{P}}_{tx}(n)$ is a set of N_F time-domain values, which is obtained by the inverse FFT of a set of values for the corresponding element of the product matrices, $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$. Each column of $\underline{\mathbf{P}}_{tx}(n)$ is a steering vector for a corresponding element of $\underline{\mathbf{s}}(n)$.

[1090] A convolver 240 receives and preconditions the modulation symbol vectors, $\underline{\mathbf{s}}(n)$, with the pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, to provide the transmitted symbol vectors, $\underline{\mathbf{x}}(n)$. In the time domain, the preconditioning is a convolution operation, and the convolution of $\underline{\mathbf{s}}(n)$ with $\underline{\mathbf{P}}_{tx}(n)$ may be expressed as:

$$\underline{\mathbf{x}}(n) = \sum_{\ell} \underline{\mathbf{P}}_{tx}(\ell)\underline{\mathbf{s}}(n-\ell) \quad . \qquad \text{Eq (13)}$$

The matrix convolution shown in equation (13) may be performed as follows. To derive the i-th element of the vector $\underline{\mathbf{x}}(n)$ for time n, $x_i(n)$, the inner product of the i-th row of the matrix $\underline{\mathbf{P}}_{tx}(\ell)$ with the vector $\underline{\mathbf{s}}(n-\ell)$ is formed for a number of delay indices (e.g., $0 \le \ell \le (N_F - 1)$), and the results are accumulated to derive the element $x_i(n)$. The preconditioned symbol streams transmitted on each transmit antenna (i.e., each element of $\underline{\mathbf{x}}(n)$ or $x_i(n)$) is thus formed as a weighted combination of the N_R modulation symbol streams, with the weighting determined by the appropriate column of the matrix $\underline{\mathbf{P}}_{tx}(n)$. The process is repeated such that each element of $\underline{\mathbf{x}}(n)$ is obtained from a respective column of the matrix $\underline{\mathbf{P}}_{tx}(n)$ and the vector $\underline{\mathbf{s}}(n)$.

[1091] Each element of $\underline{\mathbf{x}}(n)$ corresponds to a sequence of preconditioned symbols to be transmitted over a respective transmit antenna. The N_T preconditioned symbol sequences collectively form a sequence of vectors, which are also referred to as the transmitted symbol vectors, $\underline{\mathbf{x}}(n)$, with each such vector including up to N_T preconditioned symbols to be transmitted from up to N_T transmit antennas for the n-th symbol period. The N_T preconditioned symbol sequences are provided to transmitters 122a through 122t and processed to derive N_T modulated signals, which are then transmitted from antennas 124a through 124t, respectively.

[1092] The embodiment shown in FIG. 2 performs time-domain beam-steering of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. The beam-steering may also be performed in the frequency domain. This can be done using techniques, such as the overlap-add method, which are well-known in the digital signal processing field, for implementing finite-duration impulse response (FIR) filters in the frequency domain. In this case, the sequences that make up the elements of the matrix $\underline{\mathbf{P}}_{tx}(n)$ for $n = (0,\ 1,\ \ldots\ , N_F - 1)$ are each padded with $N_O - N_F$ zeros, resulting in a matrix of zero-padded sequences, $\underline{\mathbf{q}}_{tx}(n)$, for $n = (0,\ 1,\ \ldots\ , N_O - 1)$. An N_O-point fast Fourier transform (FFT) is then computed for each zero-padded sequence in the matrix $\underline{\mathbf{q}}_{tx}(n)$, resulting in a matrix $\underline{\mathbf{Q}}_{tx}(k)$ for $k = (0,\ 1,\ \ldots\ , N_O - 1)$.

[1093] Also, the sequences of modulation symbols that make up the elements of $\underline{\mathbf{s}}(n)$ are each broken up into subsequences of length $N_{SS} = N_O - N_F + 1$. Each subsequence is then padded with $N_F - 1$ zeros to provide a corresponding vector of length N_O. The sequences of $\underline{\mathbf{s}}(n)$ are thus processed to provide sequences of vectors of length N_O, $\tilde{\underline{\mathbf{s}}}_\ell(n)$, where the subscript ℓ is the index for the vectors that correspond to the zero-padded subsequences. An N_O-point fast Fourier transform is then computed for each of the zero-padded subsequences, resulting in a sequence of frequency-domain vectors, $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for different values of ℓ. Each vector $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for a given ℓ, includes a set of frequency-domain vectors of length N_O (i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$). The matrix $\underline{\mathbf{Q}}_{tx}(k)$ is then multiplied with the vector $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for each value of ℓ, where the pre-multiplication is performed for each value of k, i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$.

The inverse FFTs are then computed for the matrix-vector product $\underline{\underline{\mathbf{Q}}}_{tx}(k)\underline{\tilde{\mathbf{S}}}_{\ell}(k)$ to provide a set of time-domain subsequences of length N_O. The resulting subsequences are then reassembled, according to the overlap-add method, or similar means, as is well-known in the art, to form the desired output sequences.

[1094] **FIG. 6** is a flow diagram of an embodiment of a process 600 that may be performed at the transmitter unit to implement the various transmit processing techniques described herein. Initially, data to be transmitted (i.e., the information bits) is processed in accordance with a particular processing scheme to provide a number of streams of modulation symbols (step 612). As noted above, the processing scheme may include one or more coding schemes and one or more modulation schemes (e.g., a separate coding and modulation scheme for each modulation symbol stream).

[1095] An estimated channel response matrix for the MIMO channel is then obtained (step 614). This matrix may be the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, or the estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, which may be provided to the transmitter from the receiver. The estimated channel response matrix is then decomposed (e.g., using channel eigen-decomposition) to obtain a set of matrices of right eigen-vectors, $\underline{\underline{\mathbf{V}}}$, and a set of matrices of singular values, $\underline{\underline{\Lambda}}$ (step 616).

[1096] A number of sets of weights, $\underline{\mathbf{w}}_{ii}$, are then derived based on the matrices of singular values (step 618). One set of weight may be derived for each eigenmode used for data transmission. These weights are used to reduce or minimize intersymbol interference at the receiver by inverting the frequency response of each eigenmode selected for use.

[1097] A set of scaling values, $\underline{\mathbf{b}}$, may also be derived based on the matrices of singular values (step 620). Step 620 is optional, as indicated by the dashed box for step 620 in FIG. 6. The scaling values may be derived using water-pouring analysis and are used to adjust the transmit powers of the selected eigenmodes.

[1098] A pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, is then derived based on the matrices of right eigen-vectors, $\underline{\underline{\mathbf{V}}}$, the sets of weights, $\underline{\mathbf{w}}_{ii}$, and (if available) the set of scaling values, $\underline{\mathbf{b}}$ (step 622). The streams of modulation symbols are then preconditioned (in either the time domain or frequency domain) based on the pulse-shaping matrix to

provide a number of streams of preconditioned symbols, $\underline{\mathbf{x}}(n)$, to be transmitted over the MIMO channel (step 624).

[1099] Time-domain transmit processing with channel eigenmode decomposition and water-pouring is described in further detail in U.S. Patent Application Serial No. 10/017,038, entitled "Time-Domain Transmit and Receive Processing with Channel Eigen-mode Decomposition for MIMO Systems," filed December 7, 2001, which is assigned to the assignee of the present application and incorporated herein by reference.

[1100] **FIG. 7** is a block diagram of an embodiment of a receiver unit 700 capable of implementing various processing techniques described herein. Receiver unit 700 is an embodiment of the receiver portion of receiver system 150 in FIG. 1. Receiver unit 700 includes (1) a RX MIMO processor 160a that processes N_R received symbol streams to provide N_T recovered symbol streams and (2) a RX data processor 162a that demodulates, deinterleaves, and decodes the recovered symbols to provide decoded bits. RX MIMO processor 160a and RX data processor 162a are one embodiment of RX MIMO processor 160 and RX data processor 162, respectively, in FIG. 1.

[1101] Referring back to FIG. 1, the transmitted signals from N_T transmit antennas are received by each of N_R antennas 152a through 152r. The received signal from each antenna is routed to a respective receiver 154, which is also referred to as a front-end processor. Each receiver 154 conditions (e.g., filters, amplifies, and frequency downconverts) a respective received signal, and further digitizes the conditioned signal to provide ADC samples. Each receiver 154 may further data demodulate the ADC samples with a recovered pilot to provide a respective stream of received symbols. Receivers 154a through 154r thus provide N_R received symbol streams. These streams collectively form a sequence of vectors, which are also referred to as the received symbol vectors, $\underline{\mathbf{r}}(n)$, with each such vector including N_R received symbols from the N_R receivers 154 for the n-th symbol period. The received symbol vectors, $\underline{\mathbf{r}}(n)$, are then provided to RX MIMO processor 160a.

[1102] Within RX MIMO processor 160a, a channel estimator 712 receives the vectors $\underline{\mathbf{r}}(n)$ and derives an estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, which may be sent back to the transmitter system and used in the transmit processing. An FFT 714

then performs an FFT on the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, to obtain the estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$ (i.e., $\underline{\underline{\hat{\mathbf{H}}}} = \text{FFT}[\underline{\underline{\hat{\mathcal{H}}}}]$).

[1103] A unit 716 then performs the channel eigen-decomposition of $\underline{\hat{\mathbf{H}}}(k)$, for each frequency bin k, to obtain the corresponding matrix of left eigen-vectors, $\underline{\mathbf{U}}(k)$. Each column of $\underline{\underline{\mathbf{U}}}$, where $\underline{\underline{\mathbf{U}}} = [\underline{\mathbf{U}}(0) \;...\; \underline{\mathbf{U}}(k) \;...\; \underline{\mathbf{U}}(N_F - 1)]$, is a steering vector for a corresponding element of $\underline{\mathbf{r}}(n)$, and is used to orthogonalize the received symbol streams. An IFFT 718 then performs the inverse FFT of $\underline{\underline{\mathbf{U}}}$ to obtain a spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}(n)$, for the receiver system.

[1104] A convolver 720 then conditions the received symbol vectors, $\underline{\mathbf{r}}(n)$, with the conjugate transpose of the spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(n)$, to obtain the recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$, which are estimates of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. In the time domain, the conditioning is a convolution operation, which may be expressed as:

$$\underline{\hat{\mathbf{s}}}(n) = \sum_{\ell} \underline{\mathcal{U}}^H(\ell)\, \underline{\mathbf{r}}(n-\ell) \quad . \qquad \text{Eq (14)}$$

[1105] The pulse-shaping at the receiver may also be performed in the frequency domain, similar to that described above for the transmitter. In this case, the N_R sequences of received symbols for the N_R receive antennas, which make up the sequence of received symbol vectors, $\underline{\mathbf{r}}(n)$, are each broken up into subsequences of N_{SS} received symbols, and each subsequence is zero-padded to provide a corresponding vector of length N_O. The N_R sequences of $\underline{\mathbf{r}}(n)$ are thus processed to provide N_R sequences of vectors of length N_O, $\underline{\tilde{\mathbf{r}}}_\ell(n)$, where the subscript ℓ is the index for the vectors that correspond to the zero-padded subsequences. Each zero-padded subsequence is then transformed via an FFT, resulting in a sequence of frequency-domain vectors, $\underline{\mathbf{R}}_\ell(k)$, for different values of ℓ. Each vector $\underline{\mathbf{R}}_\ell(k)$, for a given ℓ, includes a set of frequency-domain vectors of length N_O (i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$).

[1106] The conjugate transpose of the spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(n)$, is also zero-padded and transformed via an FFT to obtain a frequency-domain

matrix, $\underline{\tilde{\mathbf{U}}}^H(k)$ for $k = (0,\ 1,\ \ldots\ , N_O - 1)$. The vector $\underline{\mathbf{R}}_\ell(k)$, for each value of ℓ, is then pre-multiplied with the conjugate transpose matrix $\underline{\tilde{\mathbf{U}}}^H(k)$ (where the pre-multiplication is performed for each value of k, i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$) to obtain a corresponding frequency-domain vector $\underline{\hat{\mathbf{S}}}_\ell(k)$. Each vector $\underline{\hat{\mathbf{S}}}_\ell(k)$, which includes a set of frequency-domain vectors of length N_O, can then be transformed via an inverse FFT to provide a corresponding set of time-domain subsequences of length N_O. The resulting subsequences are then reassembled according to the overlap-add method, or similar means, as is well-known in the art, to obtain sequences of recovered symbols, which corresponds to the set of recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$.

[1107] Thus recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$, may be characterized as a convolution in the time domain, as follows:

$$\underline{\hat{\mathbf{s}}}(n) = \sum_{\ell} \underline{\Gamma}(\ell)\,\underline{\mathbf{s}}(n-\ell) + \underline{\hat{\mathbf{z}}}(n) \quad , \qquad \text{Eq (15)}$$

where $\underline{\Gamma}(\ell)$ is the inverse FFT of $\underline{\hat{\Lambda}}(k) = \underline{\Lambda}(k)\sqrt{\underline{\mathbf{W}}(k)}$; and

$\underline{\hat{\mathbf{z}}}(n)$ is the received noise as transformed by the receiver spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(\ell)$.

The matrix $\underline{\Gamma}(n)$ is a diagonal matrix of eigen-pulses derived from the set of matrices $\underline{\underline{\hat{\Lambda}}}$, where $\underline{\underline{\hat{\Lambda}}} = [\underline{\hat{\Lambda}}(0)\ \ldots\ \underline{\hat{\Lambda}}(k)\ \ldots\ \underline{\hat{\Lambda}}(N_F - 1)]$. In particular, each diagonal element of $\underline{\Gamma}(n)$ corresponds to an eigen-pulse that is obtained as the IFFT of a set of singular values, $[\hat{\lambda}_{ii}(0)\ \ldots\ \hat{\lambda}_{ii}(k)\ \ldots\ \hat{\lambda}_{ii}(N_F - 1)]^T$, for a corresponding element of $\underline{\underline{\hat{\Lambda}}}$,

[1108] The two forms for ordering the singular values, sorted and random-ordered, result in two different types of eigen-pulses. For the sorted form, the resulting eigen-pulse matrix, $\underline{\Gamma}_s(n)$, is a diagonal matrix of pulses that are sorted in descending order of energy content. The pulse corresponding to the first diagonal element of the eigen-pulse matrix, $\{\underline{\Gamma}_s(n)\}_{11}$, has the most energy, and the pulses corresponding to elements further down the diagonal have successively less energy. Furthermore, when the SNR is low enough that water-pouring results in some of the frequency bins having little or no energy, the energy is taken out of the smallest eigen-pulses first. Thus, at low SNRs,

one or more of the eigen-pulses may have little or no energy. This has the advantage that at low SNRs, the coding and modulation are simplified through the reduction in the number of orthogonal subchannels. However, in order to approach channel capacity, the coding and modulation are performed separately for each eigen-pulse.

[1109] The random-ordered form of the singular values in the frequency domain may be used to further simplify coding and modulation (i.e., to avoid the complexity of separate coding and modulation for each element of the eigen-pulse matrix). In the random-ordered form, for each frequency bin, the ordering of the singular values is random instead of being based on their magnitude or size. This random ordering can result in approximately equal energy in all of the eigen-pulses. When the SNR is low enough to result in frequency bins with little or no energy, these bins are spread approximately evenly among the eigenmodes so that the number of eigen-pulses with non-zero energy is the same independent of SNR. At high SNRs, the random-order form has the advantage that all of the eigen-pulses have approximately equal energy, in which case separate coding and modulation for different eigenmodes are not required.

[1110] If the response of the MIMO channel is frequency selective, then the elements in the diagonal matrices, $\underline{\Lambda}(k)$, are time-dispersive. However, because of the pre-processing at the transmitter to invert the channel, the resulting recovered symbol sequences, $\underline{\hat{s}}(n)$, have little intersymbol interference, if the channel inversion is effectively performed. In that case, additional equalization would not be required at the receiver to achieve high performance.

[1111] If the channel inversion is not effective (e.g., due to an inaccurate estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$) then an equalizer may be used to equalize the recovered symbols, $\underline{\hat{s}}(n)$, prior to the demodulation and decoding. Various types of equalizer may be used to equalize the recovered symbol streams, including a minimum mean square error linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum likelihood sequence estimator (MLSE), and so on.

[1112] Since the orthogonalization process at the transmitter and receiver results in decoupled (i.e., orthogonal) recovered symbol streams at the receiver, the complexity of the equalization required for the decoupled symbol streams is greatly reduced. In particular, the equalization may be achieved by parallel time-domain equalization of the independent symbol streams. The equalization may be performed as described in the

aforementioned U.S. Patent Application Serial No. 10/017,038, and in U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001, which is assigned to the assignee of the present application and incorporated herein by reference.

[1113] For the embodiment in FIG. 7, the recovered symbol vectors, $\hat{\underline{s}}(n)$, are provided to RX data processor 162a. Within processor 162a, a symbol demapping element 732 demodulates each recovered symbol in $\hat{\underline{s}}(n)$ in accordance with a demodulation scheme that is complementary to the modulation scheme used for that symbol at the transmitter system. The demodulated data from symbol demapping element 732 is then de-interleaved by a deinterleaver 734. The deinterleaved data is further decoded by a decoder 736 to obtain the decoded bits, $\hat{d}_i$, which are estimates of the transmitted information bits, d_i. The deinterleaving and decoding are performed in a manner complementary to the interleaving and encoding, respectively, performed at the transmitter system. For example, a Turbo decoder or a Viterbi decoder may be used for decoder 736 if Turbo or convolutional coding, respectively, is performed at the transmitter system.

[1114] **FIG. 8** is a flow diagram of a process 800 that may be performed at the receiver unit to implement the various receive processing techniques described herein. Initially, an estimated channel response matrix for the MIMO channel is obtained (step 812). This matrix may be the estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, or the estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$. The matrix $\hat{\underline{\underline{\mathcal{H}}}}$ or $\hat{\underline{\underline{\mathbf{H}}}}$ may be obtained, for example, based on pilot symbols transmitted over the MIMO channel. The estimated channel response matrix is then decomposed (e.g., using channel eigen-decomposition) to obtain a set of matrices of left eigen-vectors, $\underline{\underline{\mathbf{U}}}$ (step 814).

[1115] A pulse-shaping matrix $\underline{\mathcal{U}}(n)$ is then derived based on the matrices of left eigen-vectors, $\underline{\underline{\mathbf{U}}}$ (step 816). The streams of received symbols are then conditioned (in either the time domain or frequency domain) based on the pulse-shaping matrix $\underline{\mathcal{U}}(n)$ to provide the streams of recovered symbols (step 818). The recovered symbols are further processed in accordance with a particular receive processing scheme, which is

complementary to the transmit processing scheme used at the transmitter, to provide the decoded data (step 820).

[1116] Time-domain receive processing with channel eigenmode decomposition is described in further detail in the aforementioned U.S. Patent Application Serial No. 10/017,038.

[1117] The techniques for processing a data transmission at a transmitter and a receiver described herein may be implemented in various wireless communication systems, including but not limited to MIMO and CDMA systems. These techniques may also be used for the forward link and/or the reverse link.

[1118] The techniques described herein to process a data transmission at the transmitter and receiver may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements used to perform various signal processing steps at the transmitter (e.g., to code and modulate the data, decompose the channel response matrix, derive the weights to invert the channel, derive the scaling values for power allocation, derive the transmitter pulse-shaping matrix, precondition the modulation symbols, and so on) or at the receiver (e.g., to decompose the channel response matrix, derive the receiver pulse-shaping matrix, condition the received symbols, demodulate and decode the recovered symbols, and so on) may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1119] For a software implementation, some or all of the signal processing steps at each of the transmitter and receiver may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memories 132 and 172 in FIG. 1) and executed by a processor (e.g., controllers 130 and 170). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[1120] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various

modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1121] **WHAT IS CLAIMED IS:**

CLAIMS

1. In a multiple-input multiple-output (MIMO) communication system, a method for processing data for transmission over a MIMO channel, comprising:

processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

deriving a pulse-shaping matrix based on an estimated response of the MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

2. The method of claim 1, further comprising:

deriving a plurality of weights based on an estimated channel response matrix for the MIMO channel, wherein the weights are used to invert a frequency response of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the weights.

3. The method of claim 2, further comprising:

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values, and

wherein the weights are derived based on the matrices of singular values and the pulse-shaping matrix is further derived based on the matrices of eigen-vectors.

4. The method of claim 2, wherein the estimated channel response matrix is descriptive of a plurality of eigenmodes of the MIMO channel.

5. The method of claim 4, wherein one set of weights is derived for each eigenmode used for data transmission and wherein the weights in each set are derived to invert the frequency response of the corresponding eigenmode.

6. The method of claim 4, further comprising:

deriving a plurality of scaling values based on the matrices of singular values, wherein the scaling values are used to adjust transmit powers for the eigenmodes of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

7. The method of claim 6, wherein the scaling values are derived based on water-pouring analysis.

8. The method of claim 3, wherein the estimated channel response matrix is provided in the frequency domain and is decomposed in the frequency domain.

9. The method of claim 3, wherein the estimated channel response matrix is decomposed using channel eigen-decomposition.

10. The method of claim 4, wherein eigenmodes associated with transmission capabilities below a particular threshold are not used for data transmission.

11. The method of claim 3, wherein the singular values in each matrix are sorted based on their magnitude.

12. The method of claim 4, wherein the singular values in each matrix are randomly ordered such that the eigenmodes of the MIMO channel are associated with approximately equal transmission capabilities.

13. The method of claim 1, wherein the pulse-shaping matrix comprises a plurality of sequences of time-domain values, and wherein the preconditioning is performed in the time domain by convolving the streams of modulation symbols with the pulse-shaping matrix.

14. The method of claim 1, wherein the pulse-shaping matrix comprises a plurality of sequences of frequency-domain values, and wherein the preconditioning is performed in the frequency domain by multiplying a plurality of streams of transformed modulation symbols with the pulse-shaping matrix.

15. The method of claim 1, wherein the pulse-shaping matrix is derived to maximize capacity by allocating more transmit power to eigenmodes of the MIMO channel having greater transmission capabilities.

16. The method of claim 1, wherein the pulse-shaping matrix is derived to provide approximately equal received signal-to-noise-and-interference ratios (SNRs) for the plurality of modulation symbol streams at the receiver.

17. The method of claim 1, wherein the particular processing scheme defines a separate coding and modulation scheme for each modulation symbol stream.

18. The method of claim 1, wherein the particular processing scheme defines a common coding and modulation scheme for all modulation symbol streams.

19. In a multiple-input multiple-output (MIMO) communication system, a method for processing data for transmission over a MIMO channel, comprising:

processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values;

deriving a plurality of weights based on the matrices of singular values, wherein the weights are used to invert the frequency response of the MIMO channel;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors and the weights; and

preconditioning the plurality of streams of modulation symbols based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

20. The method of claim 19, further comprising:

deriving a plurality of scaling values based on the matrices of singular values, wherein the scaling values are used to adjust transmit powers for eigenmodes of the

MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

21. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

process data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

derive a pulse-shaping matrix based on an estimated response of the MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

precondition the plurality of streams of modulation symbols based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

22. In a multiple-input multiple-output (MIMO) communication system, a method for processing a data transmission received via a MIMO channel, comprising:

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

23. The method of claim 22, wherein the conditioning is performed in the time domain based on a time-domain pulse-shaping matrix.

24. The method of claim 22, wherein the conditioning is performed in the frequency domain and includes

transforming the plurality of received symbol streams to the frequency domain;

multiplying the transformed received symbol streams with a frequency-domain pulse-shaping matrix to provide a plurality of conditioned symbol streams; and

transforming the plurality of conditioned symbol streams to the time domain to provide the plurality of recovered symbol streams.

25. The method of claim 22, wherein the conditioning orthogonalizes a plurality of streams of modulation symbols transmitted over the MIMO channel.

26. The method of claim 22, further comprising:

demodulating the plurality of recovered symbol streams in accordance with one or more demodulation schemes to provide a plurality of demodulated data streams; and

decoding the plurality of demodulated data streams in accordance with one or more decoding schemes to provide decoded data.

27. The method of claim 22, further comprising:

deriving channel state information (CSI) comprised of the estimated channel response matrix for the MIMO channel; and

sending the CSI back to the transmitter.

28. In a multiple-input multiple-output (MIMO) communication system, a method for processing a data transmission received via a MIMO channel, comprising:

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors;

conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver;

demodulating the plurality of recovered symbol streams in accordance with one or more demodulation schemes to provide a plurality of demodulated data streams; and

decoding the plurality of demodulated data streams in accordance with one or more decoding schemes to provide decoded data.

29. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

obtain an estimated channel response matrix for a MIMO channel used for a data transmission;

decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

derive a pulse-shaping matrix based on the matrices of eigen-vectors; and

condition a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

30. A transmitter unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a TX data processor operative to process data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols; and

a TX MIMO processor operative to derive a pulse-shaping matrix based on an estimated response of a MIMO channel and in a manner to reduce intersymbol interference at a receiver, and to precondition the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

31. The transmitter unit of claim 30, wherein the TX MIMO processor is further operative to derive a plurality of weights based on an estimated channel response matrix for the MIMO channel, wherein the weights are used to invert the frequency response of the MIMO channel, and wherein the pulse-shaping matrix is derived based in part on the weights.

32. The transmitter unit of claim 31, wherein the TX MIMO processor is further operative to decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values, and

wherein the weights are derived based on the matrices of singular values and the pulse-shaping matrix is further derived based on the matrices of eigen-vectors.

33. The transmitter unit of claim 31, wherein the TX MIMO processor is further operative to derive a plurality of scaling values used to adjust transmit powers for the eigenmodes of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

34. The transmitter unit of claim 33, wherein the scaling values are derived based on water-pouring analysis on a plurality of matrices of singular values obtained from the estimated channel response matrix.

35. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

means for deriving a pulse-shaping matrix based on an estimated response of a MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

means for preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

36. A digital signal processor comprising:

means for processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

means for deriving a pulse-shaping matrix based on an estimated response of a multiple-input multiple-output (MIMO) channel and in a manner to reduce intersymbol interference at a receiver; and

means for preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

37. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

an RX MIMO processor operative to obtain an estimated channel response matrix for a MIMO channel used for a data transmission, decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors, derive a pulse-shaping matrix based on the matrices of eigen-vectors, and condition a plurality of streams of received symbols based on the pulse-shaping matrix to obtain a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted over the MIMO channel, wherein the modulation symbols were preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at the receiver unit; and

an RX data processor operative to process the plurality of recovered symbol streams in accordance with a particular processing scheme to provide decoded data.

38. The receiver unit of claim 37, wherein the RX MIMO processor is operative to condition the plurality of streams of received symbols in the time domain based on a time-domain pulse-shaping matrix.

39. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for obtaining an estimated channel response matrix for a MIMO channel used for a data transmission;

means for decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

means for deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

means for conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

40. A digital signal processor comprising:

means for obtaining an estimated channel response matrix for a multiple-input multiple-output (MIMO) channel used for a data transmission;

means for decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

means for deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

means for conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

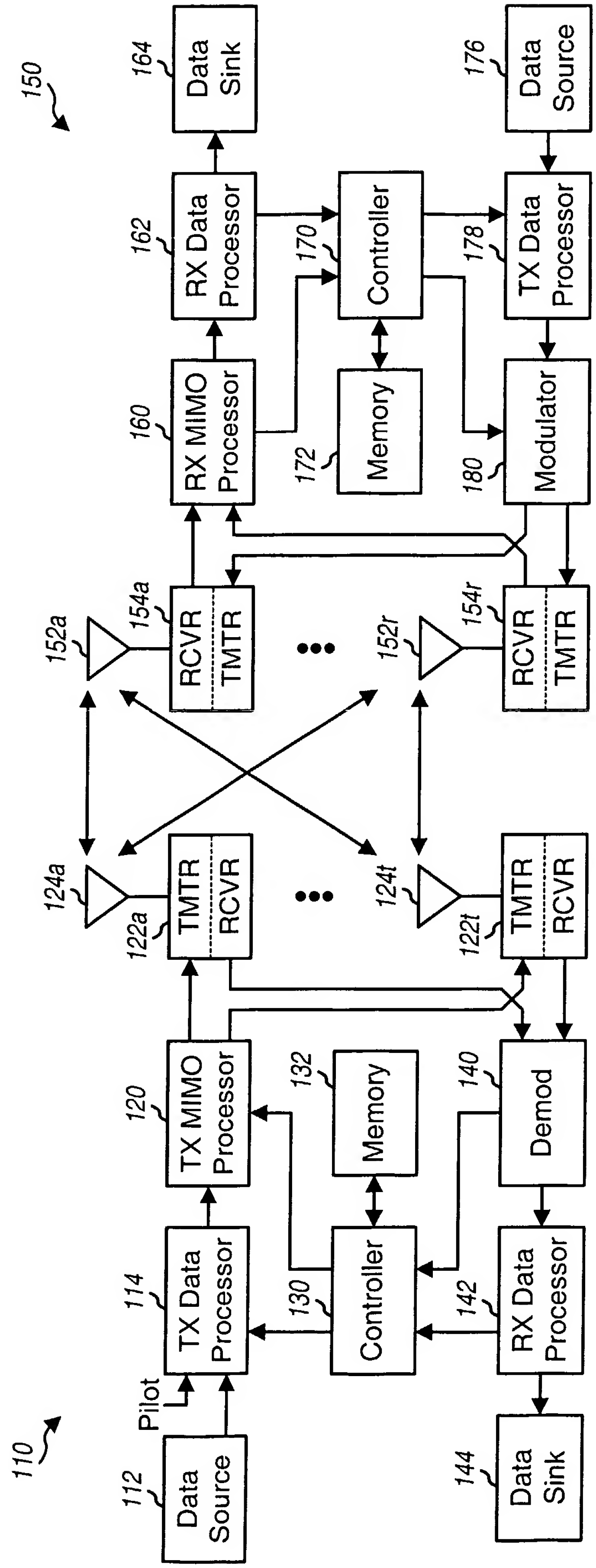

150
110
Data Source
112
Pilot
TX Data Processor
114
TX MIMO Processor
120
TMTR
RCVR
122a
124a
122t
124t
Controller
130
Memory
132
Demod
140
RX Data Processor
142
Data Sink
144
RCVR
TMTR
152a
154a
152r
154r
RX MIMO Processor
160
RX Data Processor
162
Data Sink
164
Controller
170
Memory
172
Data Source
176
TX Data Processor
178
Modulator
180

FIG. 1

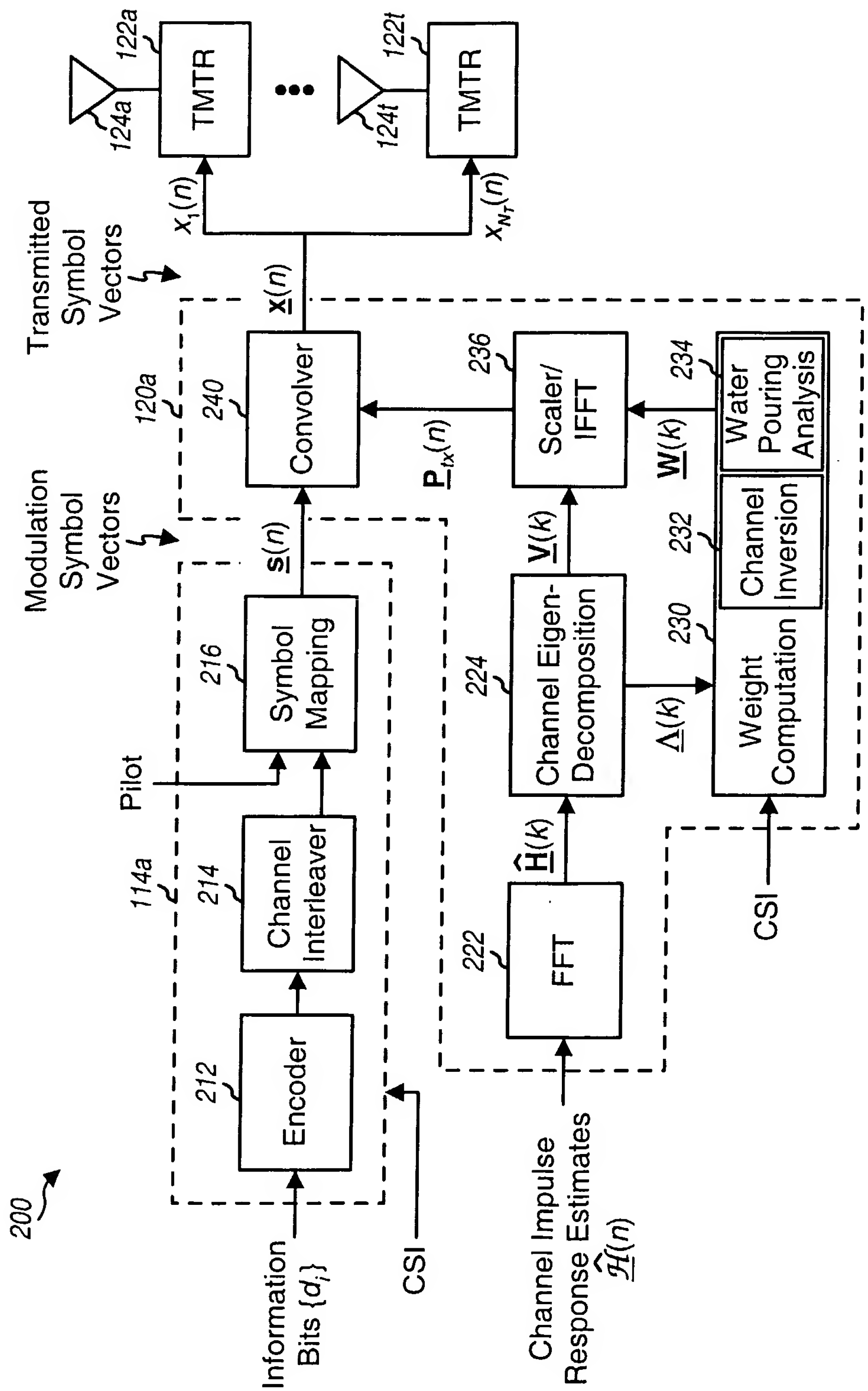

200
Information Bits $\{d_i\}$
CSI
114a
212
Encoder
214
Channel Interleaver
Pilot
216
Symbol Mapping
Modulation Symbol Vectors
$\underline{s}(n)$
120a
240
Convolver
Transmitted Symbol Vectors
$\underline{x}(n)$
$x_1(n)$
$x_{N_T}(n)$
122a
TMTR
124a
122t
TMTR
124t
Channel Impulse Response Estimates $\hat{\mathcal{H}}(n)$
222
FFT
$\hat{\underline{H}}(k)$
224
Channel Eigen-Decomposition
$\underline{V}(k)$
$\underline{\Lambda}(k)$
236
Scaler/ IFFT
$\underline{P}_{tx}(n)$
$\underline{W}(k)$
230
Weight Computation
232
Channel Inversion
234
Water Pouring Analysis
CSI

FIG. 2

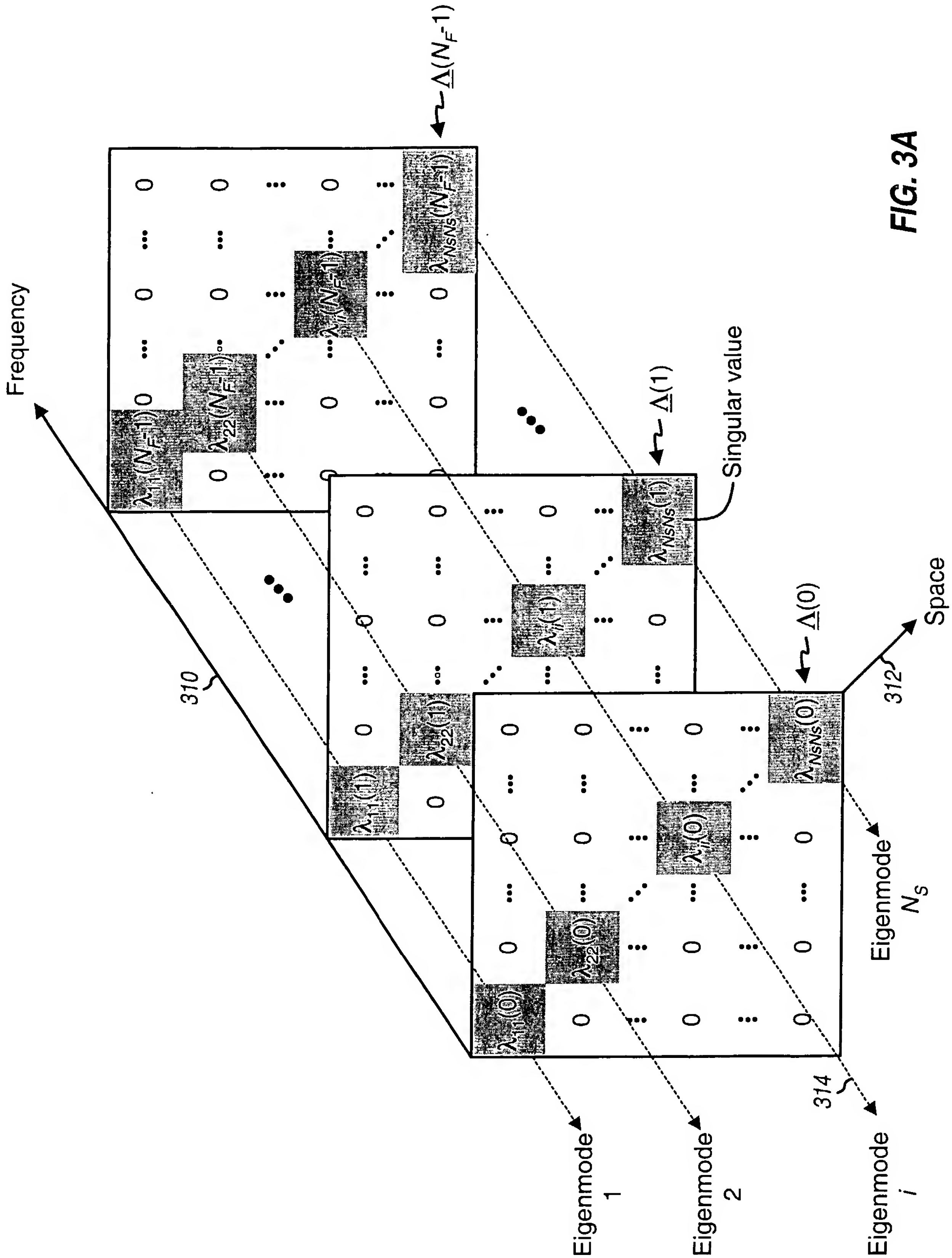

Frequency
Space
Singular value
Δ(0)
Δ(1)
Δ(N_F-1)
310
312
314
Eigenmode 1
Eigenmode 2
Eigenmode i
Eigenmode N_S

FIG. 3A

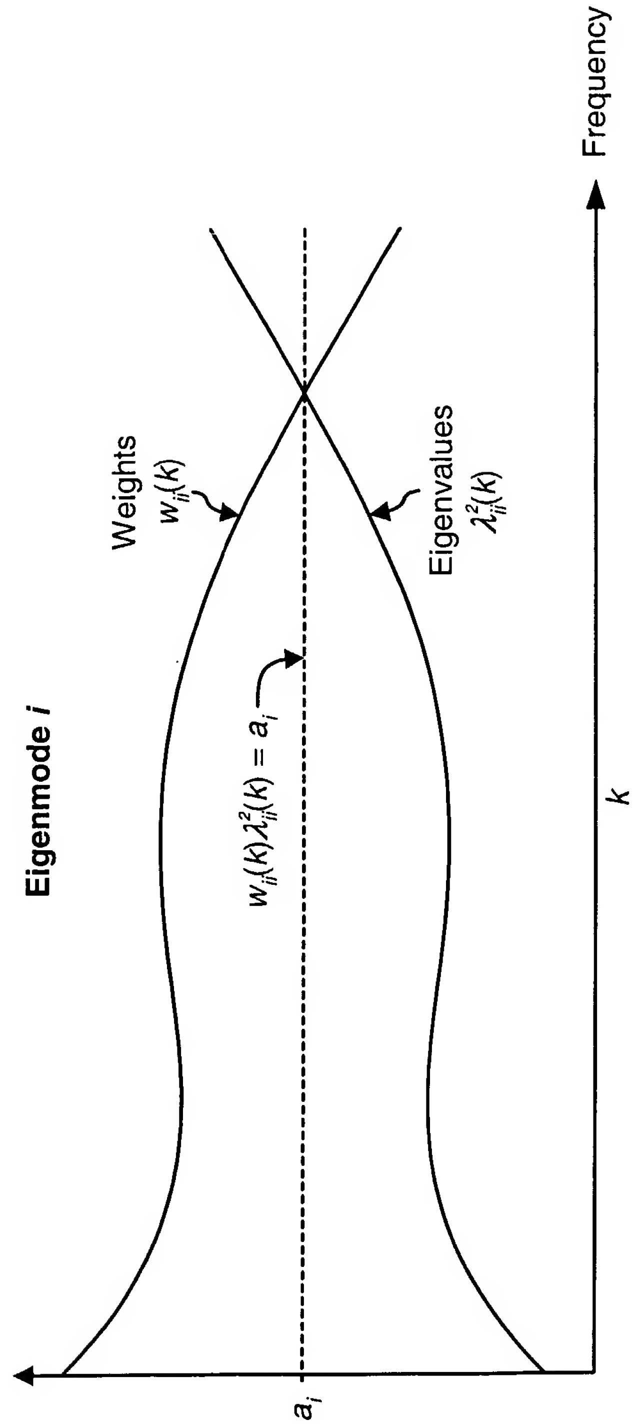

Eigenmode i
Weights $w_{ii}(k)$
Eigenvalues $\lambda_{ii}^2(k)$
$w_{ii}(k)\lambda_{ii}^2(k) = a_i$
a_i
k
Frequency

FIG. 3B

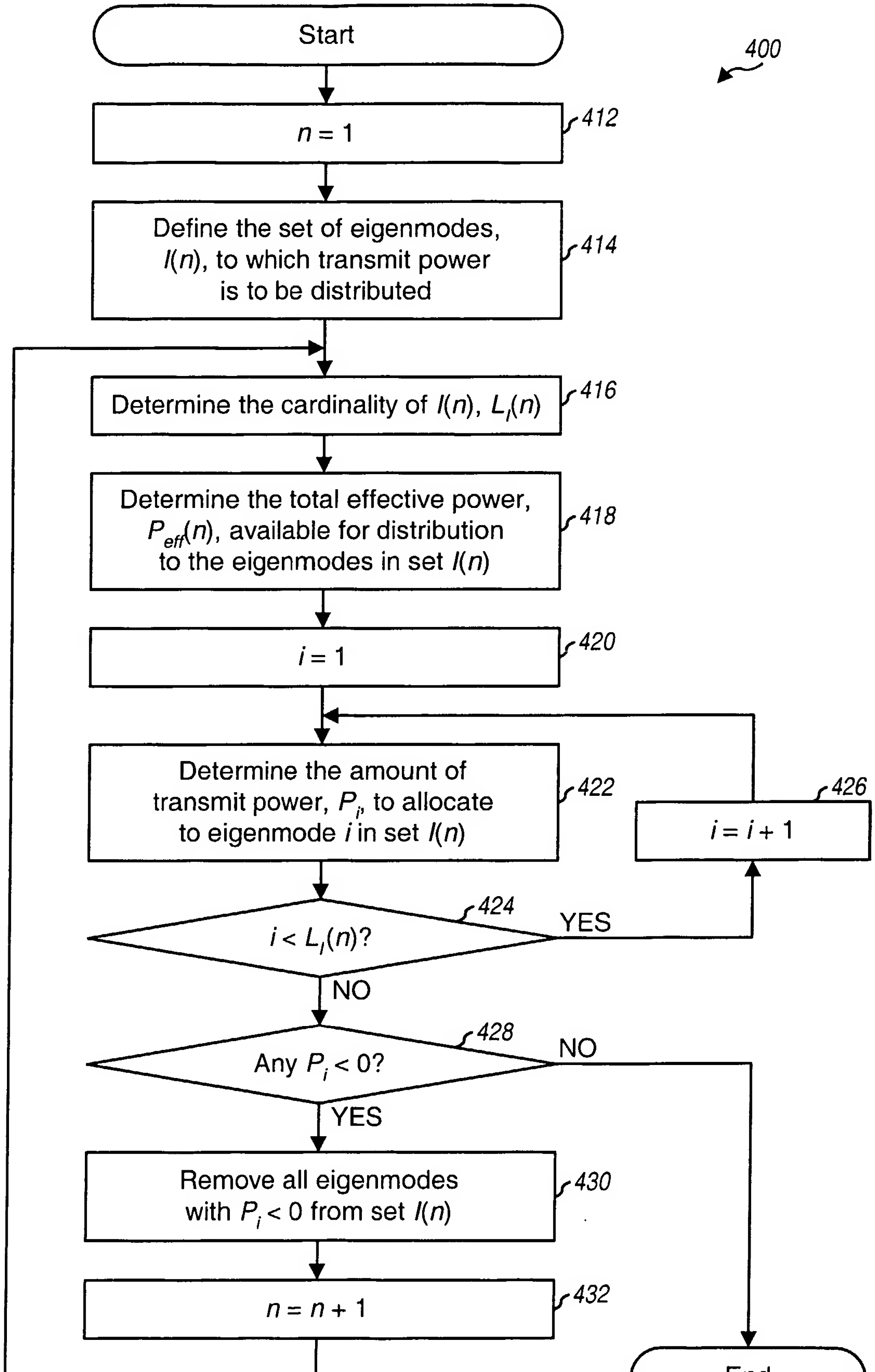

Start
400
n = 1
412
Define the set of eigenmodes, I(n), to which transmit power is to be distributed
414
Determine the cardinality of I(n), L_I(n)
416
Determine the total effective power, P_eff(n), available for distribution to the eigenmodes in set I(n)
418
i = 1
420
Determine the amount of transmit power, P_i, to allocate to eigenmode i in set I(n)
422
i = i + 1
426
i < L_I(n)?
424
YES
NO
Any P_i < 0?
428
NO
YES
Remove all eigenmodes with P_i < 0 from set I(n)
430
n = n + 1
432
End

FIG. 4

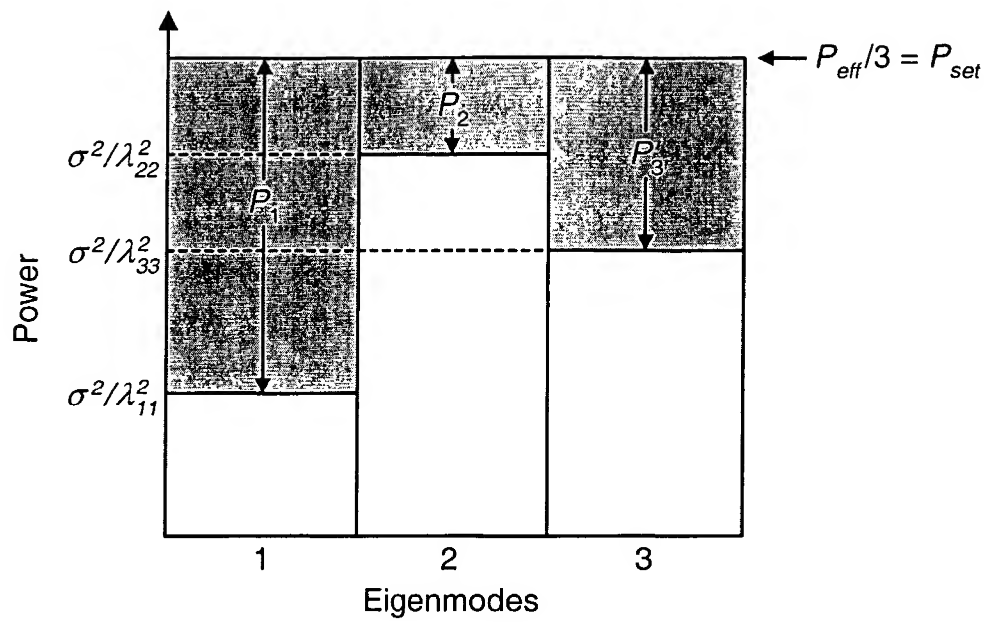

$P_{eff}/3 = P_{set}$
σ^2/λ_{22}^2
σ^2/λ_{33}^2
σ^2/λ_{11}^2
P_1
P_2
P_3
Power
1
2
3
Eigenmodes

FIG. 5A

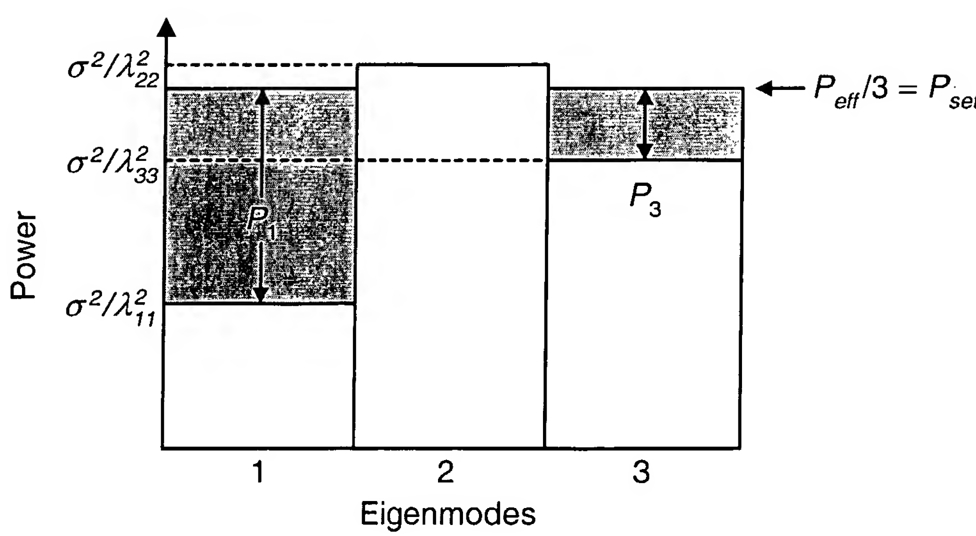

σ^2/λ_{22}^2
$P_{eff}/3 = P_{set}$
σ^2/λ_{33}^2
σ^2/λ_{11}^2
P_1
P_3
Power
1
2
3
Eigenmodes

FIG. 5B

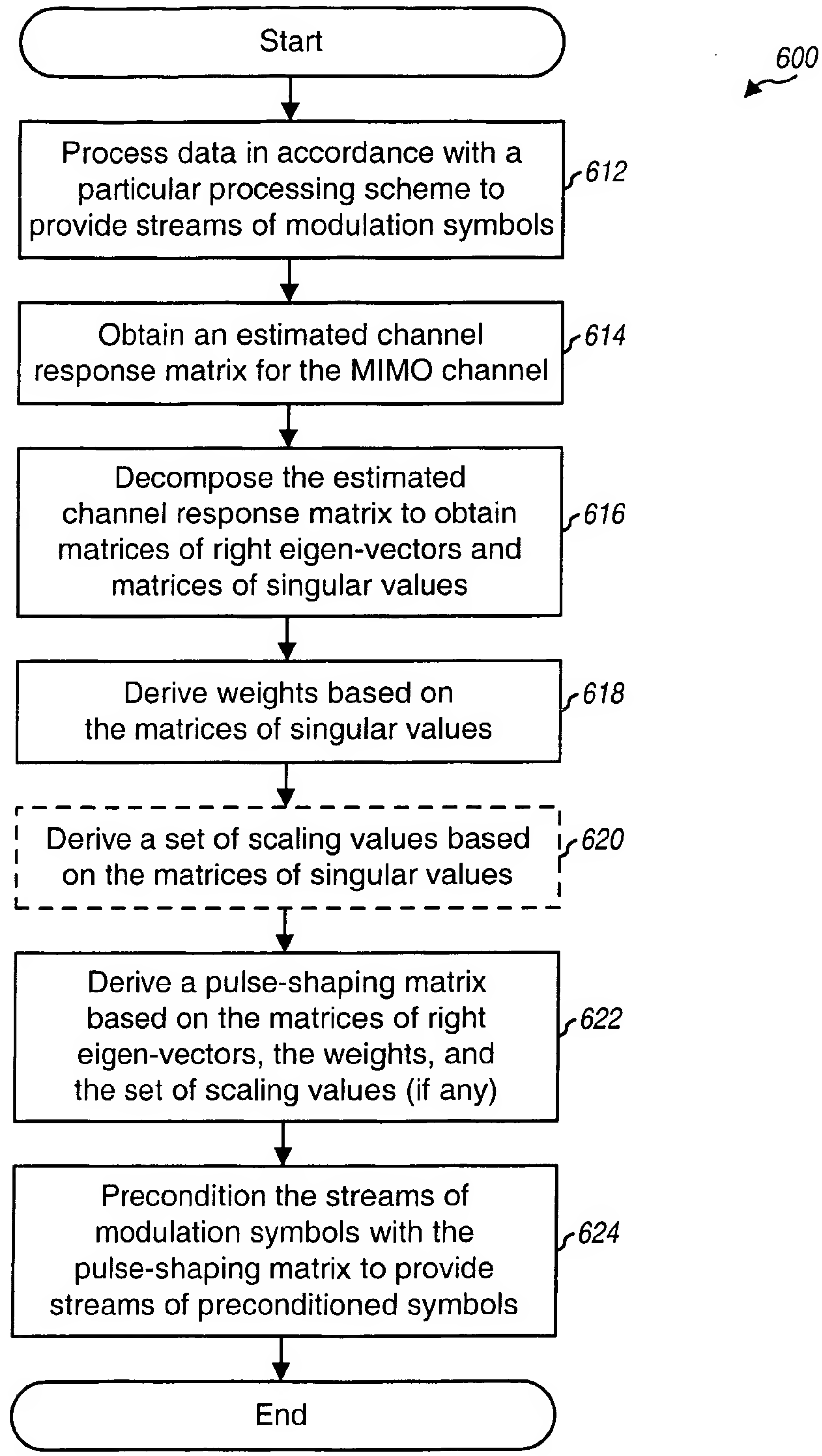

600
Start
Process data in accordance with a particular processing scheme to provide streams of modulation symbols
612
Obtain an estimated channel response matrix for the MIMO channel
614
Decompose the estimated channel response matrix to obtain matrices of right eigen-vectors and matrices of singular values
616
Derive weights based on the matrices of singular values
618
Derive a set of scaling values based on the matrices of singular values
620
Derive a pulse-shaping matrix based on the matrices of right eigen-vectors, the weights, and the set of scaling values (if any)
622
Precondition the streams of modulation symbols with the pulse-shaping matrix to provide streams of preconditioned symbols
624
End

FIG. 6

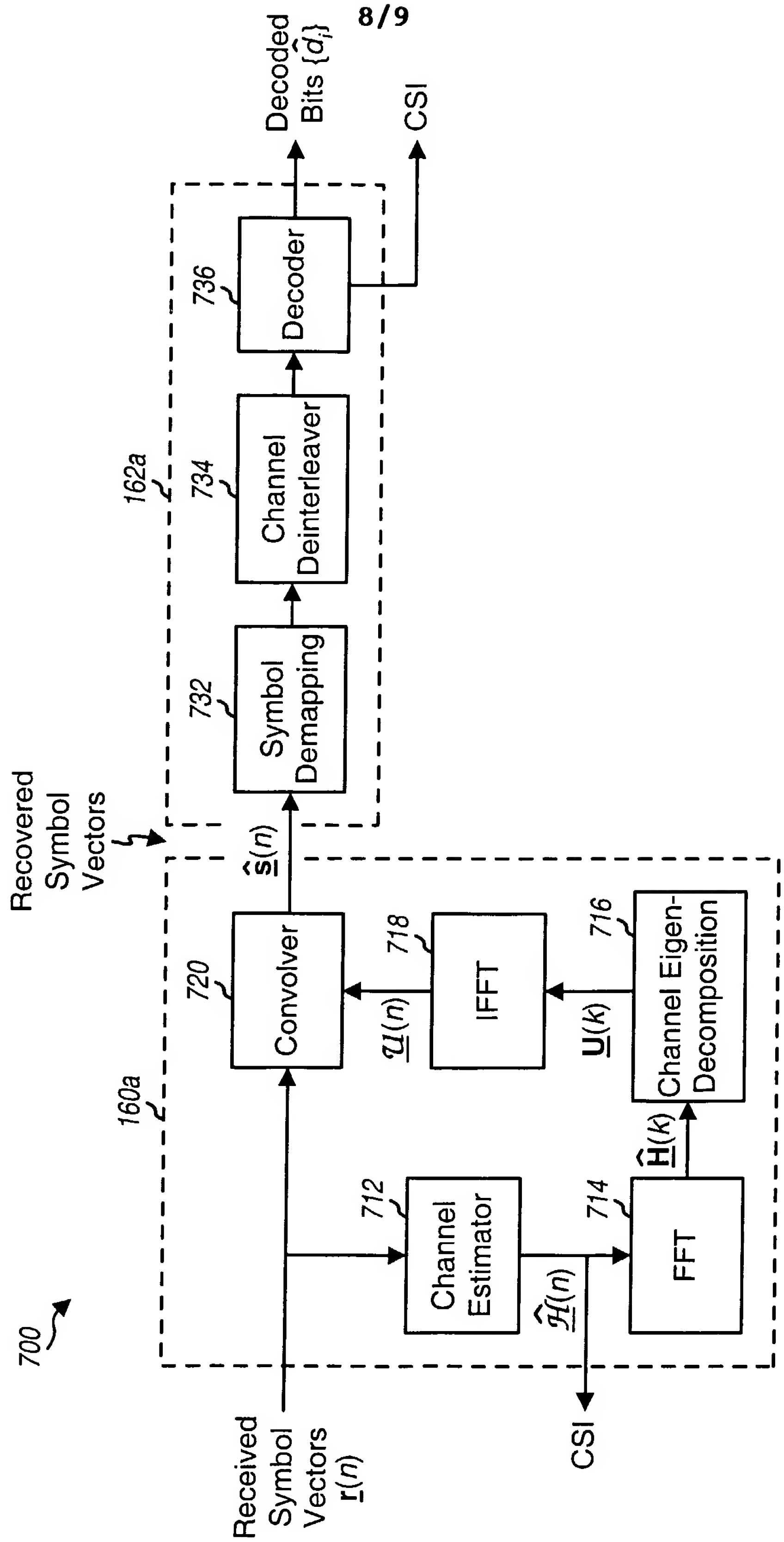

700
Received Symbol Vectors $\underline{\mathbf{r}}(n)$
160a
712
Channel Estimator
$\hat{\underline{\mathcal{H}}}(n)$
CSI
714
FFT
$\hat{\underline{\mathbf{H}}}(k)$
716
Channel Eigen-Decomposition
$\underline{\mathbf{U}}(k)$
718
IFFT
$\underline{\mathcal{U}}(n)$
720
Convolver
$\hat{\underline{\mathbf{s}}}(n)$
Recovered Symbol Vectors
162a
732
Symbol Demapping
734
Channel Deinterleaver
736
Decoder
Decoded Bits $\{\hat{d}_i\}$
CSI

FIG. 7

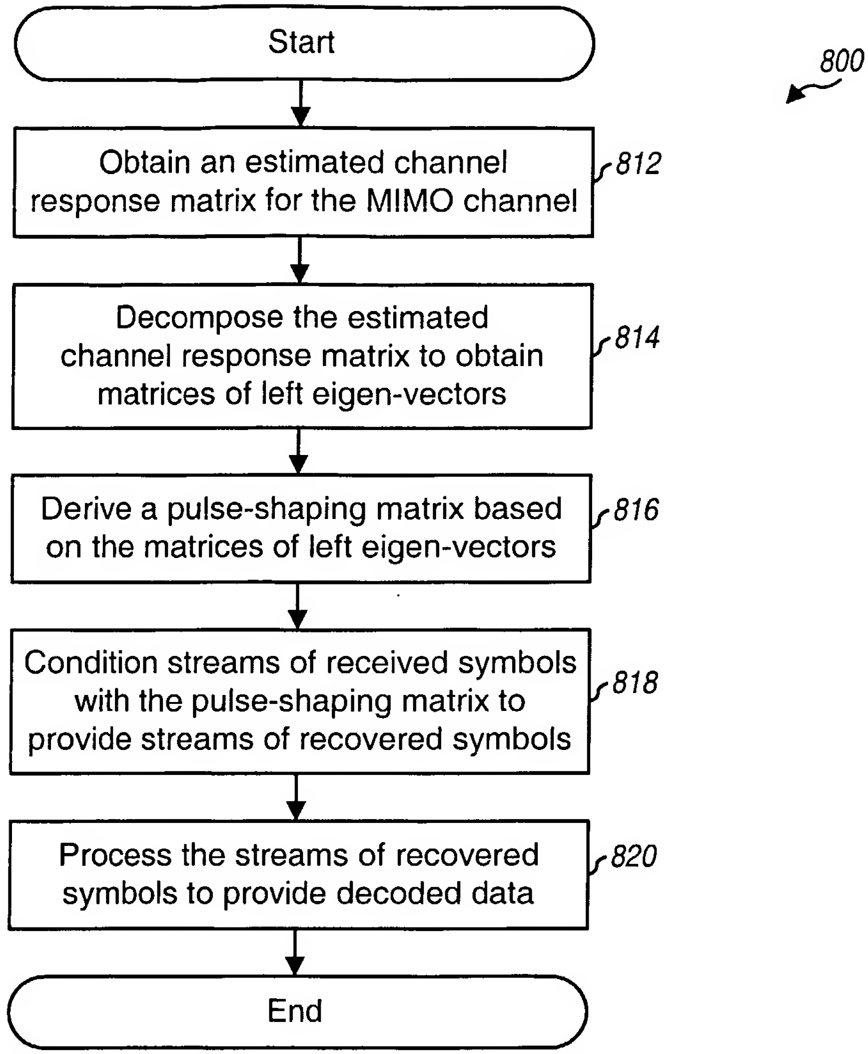

Start
Obtain an estimated channel response matrix for the MIMO channel
812
Decompose the estimated channel response matrix to obtain matrices of left eigen-vectors
814
Derive a pulse-shaping matrix based on the matrices of left eigen-vectors
816
Condition streams of received symbols with the pulse-shaping matrix to provide streams of recovered symbols
818
Process the streams of recovered symbols to provide decoded data
820
End
800

FIG. 8

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 03/19464

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/06 H04L25/03 H04L25/02

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	SAMPATH H ET AL: "Joint transmit and receive optimization for high data rate wireless communication using multiple antennas" SIGNALS, SYSTEMS, AND COMPUTERS, 1999. CONFERENCE RECORD OF THE THIRTY-THIRD ASILOMAR CONFERENCE ON OCT. 24-27, 1999, PISCATAWAY, NJ, USA,IEEE, US, 24 October 1999 (1999-10-24), pages 215-219, XP010373976 ISBN: 0-7803-5700-0 abstract page 215, paragraph 2 page 215, paragraph 5 -page 216, paragraph 2 page 217, column 2 -page 218, paragraph 1	1,13-18, 21-30, 35-40
A	---	2-12,19, 20,31-34

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[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
24 September 2003	06/10/2003

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Reilly, D

Form PCT/ISA/210 (second sheet) (July 1992)

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	WO 02 05506 A (QUALCOMM INC) 17 January 2002 (2002-01-17) abstract page 13, line 4 - line 11 page 14, line 24 -page 19, line 4	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34
X	JOONSUK KIM ET AL: "Transmission Optimization with a Space-Time Filter at Low SNR Wireless Environment" GLOBECOM'99, vol. 1B, 5 December 1999 (1999-12-05), pages 889-893, XP010373673 abstract page 890, column 1, last paragraph -page 892, column 1, paragraph 2	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34
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A		2-12,19, 20,31-34
A	EP 0 549 019 A (KONINKL PHILIPS ELECTRONICS NV) 30 June 1993 (1993-06-30) abstract page 2, line 25 - line 30	2-12,19, 20,31-34

INTERNATIONAL SEARCH REPORT

Information on patent family members

Internatic pplication No
PCT/US 03/19464

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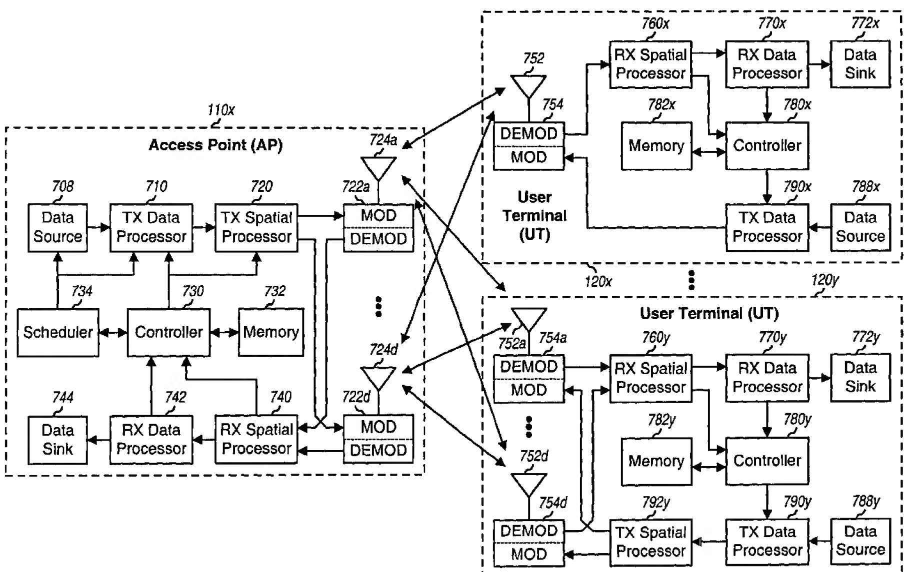
(54) Title: MIMO WLAN SYSTEM



(57) Abstract: A multiple-access MIMO WLAN system that employs MIMO, OFDM, and TDD. The system (1) uses a channel structure with a number of configurable transport channels, (2) supports multiple rates and transmission modes, which are configurable based on channel conditions and user terminal capabilities, (3) employs a pilot structure with several types of pilot (e.g., beacon, MIMO, steered reference, and carrier pilots) for different functions, (4) implements rate, timing, and power control loops for proper system operation, and (5) employs random access for system access by the user terminals, fast acknowledgment, and quick resource assignments. Calibration may be performed to account for differences in the frequency responses of transmit/receive chains at the access point and user terminals. The spatial processing may then be simplified by taking advantage of the reciprocal nature of the downlink and uplink and the calibration.

WO 2004/039011 A2

MIMO WLAN SYSTEM

Claim of Priority under 35 U.S.C. §119

[0001] This application claims the benefit of provisional U.S. Application Serial No. 60/421,309, entitled, "MIMO WLAN System," filed on October 25, 2002.

BACKGROUND

Field

[0002] The present invention relates generally to data communication, and more specifically to a multiple-input multiple-output (MIMO) wireless local area network (WLAN) communication system.

Background

[0003] Wireless communication systems are widely deployed to provide various types of communication such as voice, packet data, and so on. These systems may be multiple-access systems capable of supporting communication with multiple users sequentially or simultaneously by sharing the available system resources. Examples of multiple-access systems include Code Division Multiple Access (CDMA) systems, Time Division Multiple Access (TDMA) systems, and Frequency Division Multiple Access (FDMA) systems.

[0004] Wireless local area networks (WLANs) are also widely deployed to enable communication among wireless electronic devices (e.g., computers) via wireless link. A WLAN may employ access points (or base stations) that act like hubs and provide connectivity for the wireless devices. The access points may also connect (or "bridge") the WLAN to wired LANs, thus allowing the wireless devices access to LAN resources.

[0005] In a wireless communication system, a radio frequency (RF) modulated signal from a transmitter unit may reach a receiver unit via a number of propagation paths. The characteristics of the propagation paths typically vary over time due to a number of factors, such as fading and multipath. To provide diversity against deleterious path effects and improve performance, multiple transmit and receive antennas may be used. If the propagation paths between the transmit and receive antennas are linearly independent (i.e., a transmission on one path is not formed as a linear combination of

the transmissions on the other paths), which is generally true to at least an extent, then the likelihood of correctly receiving a data transmission increases as the number of antennas increases. Generally, diversity increases and performance improves as the number of transmit and receive antennas increases.

[0006] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S spatial channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S spatial channels corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity and/or greater reliability) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[0007] The resources for a given communication system are typically limited by various regulatory constraints and requirements and by other practical considerations. However, the system may be required to support a number of terminals, provide various services, achieve certain performance goals, and so on.

[0008] There is, therefore, a need in the art for a MIMO WLAN system capable of supporting multiple users and providing high system performance.

SUMMARY

[0009] A multiple-access MIMO WLAN system having various capabilities and able to achieve high performance is described herein. In an embodiment, the system employs MIMO and orthogonal frequency division multiplexing (OFDM) to attain high throughput, combat deleterious path effects, and provide other benefits. Each access point in the system can support multiple user terminals. The allocation of downlink and uplink resources is dependent on the requirements of the user terminals, the channel conditions, and other factors.

[0010] A channel structure supporting efficient downlink and uplink transmissions is also provided herein. The channel structure comprises a number of transport channels that may be used for a number of functions, such as signaling of system parameters and resource assignments, downlink and uplink data transmissions, random access of the system, and so on. Various attributes of these transport channels are configurable, which allows the system to easily adapt to changing channel and loading conditions.

[0011] Multiple rates and transmission modes are supported by the MIMO WLAN system to attain high throughput when supported by the channel conditions and the capabilities of the user terminals. The rates are configurable based on estimates of the channel conditions and may be independently selected for the downlink and uplink. Different transmission modes may also be used, depending on the number of antennas at the user terminals and the channel conditions. Each transmission mode is associated with different spatial processing at the transmitter and receiver and may be selected for use under different operating conditions. The spatial processing facilitates data transmission from multiple transmit antennas and/or data reception with multiple receive antennas for higher throughput and/or diversity.

[0012] In an embodiment, the MIMO WLAN system uses a single frequency band for both the downlink and uplink, which share the same operating band using time division duplexing (TDD). For a TDD system, the downlink and uplink channel responses are reciprocal. Calibration techniques are provided herein to determine and account for differences in the frequency responses of the transmit/receive chains at the access point and user terminals. Techniques are also described herein to simplify the spatial processing at the access point and user terminals by taking advantage of the reciprocal nature of the downlink and uplink and the calibration.

[0013] A pilot structure with several types of pilot used for different functions is also provided. For example, a beacon pilot may be used for frequency and system acquisition, a MIMO pilot may be used for channel estimation, a steered reference (i.e., a steered pilot) may be used for improved channel estimation, and a carrier pilot may be used for phase tracking.

[0014] Various control loops for proper system operation are also provided. Rate control may be exercised independently on the downlink and uplink. Power control may be exercised for certain transmissions (e.g., fixed-rate services). Timing control may be used for uplink transmissions to account for different propagation delays of user terminals located throughout the system.

[0015] Random access techniques to allow user terminals to access the system are also provided. These techniques support system access by multiple user terminals, fast acknowledgment of system access attempts, and quick assignment of downlink/uplink resources.

[0016] The various aspects and embodiments of the invention are described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[0017] The features and nature of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[0018] FIG. 1 shows a MIMO WLAN system;

[0019] FIG. 2 shows a layer structure for the MIMO WLAN system;

[0020] FIGS. 3A, 3B and 3C show a TDD-TDM frame structure, an FDD-TDM frame structure, and an FDD-CDM frame structure, respectively;

[0021] FIG. 4 shows the TDD-TDM frame structure with five transport channels - BCH, FCCH, FCH, RCH, and RACH;

[0022] FIGS. 5A through 5G show various protocol data unit (PDU) formats for the five transport channels;

[0023] FIG. 6 shows a structure for an FCH/RCH packet;

[0024] FIG. 7 shows an access point and two user terminals;

[0025] FIGS. 8A, 9A, and 10A show three transmitter units for the diversity, spatial multiplexing, and beam-steering modes, respectively;

[0026] FIGS. 8B, 9B, and 10B show three TX diversity processors for the diversity, spatial multiplexing, and beam-steering modes, respectively,

[0027] FIG. 8C shows an OFDM modulator;

[0028] FIG. 8D shows an OFDM symbol;

[0029] FIG. 11A shows a framing unit and a scrambler within a TX data processor;

[0030] FIG. 11B shows an encoder and a repeat/puncture unit within the TX data processor;

[0031] FIG. 11C shows another TX data processor that may be used for the spatial multiplexing mode;

[0032] FIGS. 12A and 12B show a state diagram for operation of a user terminal;

[0033] FIG. 13 shows a timeline for the RACH;

[0034] FIGS. 14A and 14B show processes for controlling the rates of downlink and uplink transmissions, respectively;

[0035] FIG. 15 shows the operation of a power control loop; and

[0036] FIG. 16 shows a process for adjusting the uplink timing of a user terminal.

DETAILED DESCRIPTION

[0037] The word "exemplary" is used herein to mean "serving as an example, instance, or illustration." Any embodiment or design described herein as "exemplary" is not necessarily to be construed as preferred or advantageous over other embodiments or designs.

I. Overall System

[0038] **FIG. 1** shows a MIMO WLAN system 100 that supports a number of users and is capable of implementing various aspects and embodiments of the invention. MIMO WLAN system 100 includes a number of access points (APs) 110 that support communication for a number of user terminals (UTs) 120. For simplicity, only two access points 110 are shown in FIG. 1. An access point is generally a fixed station that is used for communicating with the user terminals. An access point may also be referred to as a base station or some other terminology.

[0039] User terminals 120 may be dispersed throughout the system. Each user terminal may be a fixed or mobile terminal that can communicate with the access point. A user terminal may also be referred to as a mobile station, a remote station, an access terminal, a user equipment (UE), a wireless device, or some other terminology. Each user terminal may communicate with one or possibly multiple access points on the downlink and/or uplink at any given moment. The downlink (i.e., forward link) refers to transmission from the access point to the user terminal, and the uplink (i.e., reverse link) refers to transmission from the user terminal to the access point.

[0040] In FIG. 1, access point 110a communicates with user terminals 120a through 120f, and access point 110b communicates with user terminals 120f through 120k. Depending on the specific design of system 100, an access point may communicate with multiple user terminals simultaneously (e.g., via multiple code channels or subbands) or sequentially (e.g., via multiple time slots). At any given moment, a user terminal may receive downlink transmissions from one or multiple access points. The downlink transmission from each access point may include overhead data intended to be received by multiple user terminals, user-specific data intended to be received by specific user

terminals, other types of data, or any combination thereof. The overhead data may include pilot, page and broadcast messages, system parameters, and so on.

[0041] The MIMO WLAN system is based on a centralized controller network architecture. Thus, a system controller 130 couples to access points 110 and may further couple to other systems and networks. For example, system controller 130 may couple to a packet data network (PDN), a wired local area network (LAN), a wide area network (WAN), the Internet, a public switched telephone network (PSTN), a cellular communication network, and so on. System controller 130 may be designed to perform a number of functions such as (1) coordination and control for the access points coupled to it, (2) routing of data among these access points, (3) access and control of communication with the user terminals served by these access points, and so on.

[0042] The MIMO WLAN system may be able to provide high throughput with significantly greater coverage capability than conventional WLAN systems. The MIMO WLAN system can support synchronous, asynchronous, and isochronous data/voice services. The MIMO WLAN system may be designed to provide the following features:

- High service reliability
- Guaranteed quality of service (QoS)
- High instantaneous data rates
- High spectral efficiency
- Extended coverage range.

[0043] The MIMO WLAN system may be operated in various frequency bands (e.g., the 2.4 GHz and 5.x GHz U-NII bands), subject to the bandwidth and emission constraints specific to the selected operating band. The system is designed to support both indoor and outdoor deployments, with typical maximum cell size of 1 km or less. The system supports fixed terminal applications, although some operating modes also support portable and limited mobility operation.

1. MIMO, MISO, and SIMO

[0044] In a specific embodiment and as described throughout the specification, each access point is equipped with four transmit and receive antennas for data transmission and reception, where the same four antennas are used to transmit and to receive. The system also supports the case where the transmit and receive antennas of the device (e.g. access point, user terminal) are not shared, even though this configuration normally

provides lower performance than when the antennas are shared. The MIMO WLAN system may also be designed such that each access point is equipped with some other number of transmit/receive antennas. Each user terminal may be equipped with a single transmit/receive antenna or multiple transmit/receive antennas for data transmission and reception. The number of antennas employed by each user terminal type may be dependent on various factors such as, for example, the services to be supported by the user terminal (e.g., voice, data, or both), cost considerations, regulatory constraints, safety issues, and so on.

[0045] For a given pairing of multi-antenna access point and multi-antenna user terminal, a MIMO channel is formed by the N_T transmit antennas and N_R receive antennas available for use for data transmission. Different MIMO channels are formed between the access point and different multi-antenna user terminals. Each MIMO channel may be decomposed into N_S spatial channels, with $N_S \leq \min\{N_T, N_R\}$. N_S data streams may be transmitted on the N_S spatial channels. Spatial processing is required at a receiver and may or may not be performed at a transmitter in order to transmit multiple data streams on the N_S spatial channels.

[0046] The N_S spatial channels may or may not be orthogonal to one another. This depends on various factors such as (1) whether or not spatial processing was performed at the transmitter to obtain orthogonal spatial channels and (2) whether or not the spatial processing at both the transmitter and the receiver was successful in orthogonalizing the spatial channels. If no spatial processing is performed at the transmitter, then the N_S spatial channels may be formed with N_S transmit antennas and are unlikely to be orthogonal to one another.

[0047] The N_S spatial channels may be orthogonalized by performing decomposition on a channel response matrix for the MIMO channel, as described below. Each spatial channel is referred to as an eigenmode of the MIMO channel if the N_S spatial channels are orthogonalized using decomposition, which requires spatial processing at both the transmitter and the receiver, as described below. In this case, N_S data streams may be transmitted orthogonally on the N_S eigenmodes. However, an eigenmode normally refers to a theoretical construct. The N_S spatial channels are typically not completely orthogonal to one another due to various reasons. For example, the spatial channels would not be orthogonal if (1) the transmitter has no knowledge of the MIMO channel or (2) the transmitter and/or receiver have imperfect estimate of the MIMO channel.

For simplicity, in the following description, the term "eigenmode" is used to denote the case where an attempt is made to orthogonalize the spatial channels using decomposition, even though the attempt may not be fully successful due to, for example, an imperfect channel estimate.

[0048] For a given number of (e.g., four) antennas at the access point, the number of spatial channels available for each user terminal is dependent on the number of antennas employed by that user terminal and the characteristics of the wireless MIMO channel that couples the access point antennas and the user terminal antennas. If a user terminal is equipped with one antenna, then the four antennas at the access point and the single antenna at the user terminal form a multiple-input single-output (MISO) channel for the downlink and a single-input multiple-output (SIMO) channel for the uplink.

[0049] The MIMO WLAN system may be designed to support a number of transmission modes. Table 1 lists the transmission modes supported by an exemplary design of the MIMO WLAN system.

Table 1

Transmission modes	Description
SIMO	Data is transmitted from a single antenna but may be received by multiple antennas for receive diversity.
Diversity	Data is redundantly transmitted from multiple transmit antennas and/or multiple subbands to provide diversity.
Beam-steering	Data is transmitted on a single (best) spatial channel at full power using phase steering information for the principal eigenmode of the MIMO channel.
Spatial multiplexing	Data is transmitted on multiple spatial channels to achieve higher spectral efficiency.

For simplicity, the term "diversity" refers to transmit diversity in the following description, unless noted otherwise.

[0050] The transmission modes available for use for the downlink and uplink for each user terminal are dependent on the number of antennas employed at the user terminal. Table 2 lists the transmission modes available for different terminal types for the downlink and uplink, assuming multiple (e.g., four) antennas at the access point.

Table 2

Transmission modes	Downlink		Uplink	
	Single-antenna user terminal	Multi-antenna user terminal	Single-antenna user terminal	Multi-antenna user terminal
MISO (on downlink)/ SIMO (on uplink)	X	X	X	X
Diversity	X	X		X
Beam-steering	X	X		X
Spatial multiplexing		X		X

For the downlink, all of the transmission modes except for the spatial multiplexing mode may be used for single-antenna user terminals, and all transmission modes may be used for multi-antenna user terminals. For the uplink, all transmission modes may be used by multi-antenna user terminals, while single-antenna user terminals use the SIMO mode to transmit data from the one available antenna. Receive diversity (i.e., receiving a data transmission with multiple receive antennas) may be used for the SIMO, diversity, and beam-steering modes.

[0051] The MIMO WLAN system may also be designed to support various other transmission modes, and this is within the scope of the invention. For example, a beam-forming mode may be used to transmit data on a single eigenmode using both the amplitude and phase information for the eigenmode (instead of only the phase information, which is all that is used by the beam-steering mode). As another example, a "non-steered" spatial multiplexing mode can be defined whereby the transmitter simply transmits multiple data streams from multiple transmit antennas (without any spatial processing) and the receiver performs the necessary spatial processing to isolate and recover the data streams sent from the multiple transmit antennas. As yet another example, a "multi-user" spatial multiplexing mode can be defined whereby the access point transmits multiple data streams from multiple transmit antennas (with spatial processing) to multiple user terminals concurrently on the downlink. As yet another example, a spatial multiplexing mode can be defined whereby the transmitter performs spatial processing to attempt to orthogonalize the multiple data streams sent on the multiple transmit antennas (but may not be completely successful because of an

imperfect channel estimate) and the receiver performs the necessary spatial processing to isolate and recover the data streams sent from the multiple transmit antennas. Thus, the spatial processing to transmit multiple data streams via multiple spatial channels may be performed (1) at both the transmitter and receiver, (2) at only the receiver, or (3) at only the transmitter. Different spatial multiplexing modes may be used depending on, for example, the capabilities of the access point and the user terminals, the available channel state information, system requirements, and so on.

[0052] In general, the access points and user terminals may be designed with any number of transmit and receive antennas. For clarity, specific embodiments and designs are described below whereby each access point is equipped with four transmit/receive antennas, and each user terminal is equipped with four or less transmit/receive antennas.

2. OFDM

[0053] In an embodiment, the MIMO WLAN system employs OFDM to effectively partition the overall system bandwidth into a number of (N_F) orthogonal subbands. These subbands are also referred to as tones, bins, or frequency channels. With OFDM, each subband is associated with a respective subcarrier that may be modulated with data. For a MIMO system that utilizes OFDM, each spatial channel of each subband may be viewed as an independent transmission channel where the complex gain associated with each subband is effectively constant across the subband bandwidth.

[0054] In an embodiment, the system bandwidth is partitioned into 64 orthogonal subbands (i.e., $N_F = 64$), which are assigned indices of –32 to +31. Of these 64 subbands, 48 subbands (e.g., with indices of $\pm\{1, \ldots, 6, 8, \ldots, 20, 22, \ldots, 26\}$) are used for data, 4 subbands (e.g., with indices of $\pm\{7, 21\}$) are used for pilot and possibly signaling, the DC subband (with index of 0) is not used, and the remaining subbands are also not used and serve as guard subbands. This OFDM subband structure is described in further detail in a document for IEEE Standard 802.11a and entitled "Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) Specifications: High-speed Physical Layer in the 5 GHz Band," September 1999, which is publicly available and incorporated herein by reference. Different numbers of subbands and various other OFDM subband structures may also be implemented for the MIMO WLAN system, and this is within the scope of the invention. For example, all 53 subbands with indices from -26 to +26 may be used for data transmission. As another example, a 128-subband structure, a 256-subband structure, or a subband

structure with some other number of subbands may be used. For clarity, the MIMO WLAN system is described below with the 64-subband structure described above.

[0055] For OFDM, the data to be transmitted on each subband is first modulated (i.e., symbol mapped) using a particular modulation scheme selected for use for that subband. Zeros are provided for the unused subbands. For each symbol period, the modulation symbols and zeros for all N_F subbands are transformed to the time domain using an inverse fast Fourier transform (IFFT) to obtain a transformed symbol that contains N_F time-domain samples. The duration of each transformed symbol is inversely related to the bandwidth of each subband. In one specific design for the MIMO WLAN system, the system bandwidth is 20 MHz, $N_F = 64$, the bandwidth of each subband is 312.5 KHz, and the duration of each transformed symbol is 3.2 μsec.

[0056] OFDM can provide certain advantages, such as the ability to combat frequency selective fading, which is characterized by different channel gains at different frequencies of the overall system bandwidth. It is well known that frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. The ISI distortion degrades performance by impacting the ability to correctly detect the received symbols. Frequency selective fading can be conveniently combated with OFDM by repeating a portion of (or appending a cyclic prefix to) each transformed symbol to form a corresponding OFDM symbol, which is then transmitted.

[0057] The length of the cyclic prefix (i.e., the amount to repeat) for each OFDM symbol is dependent on the delay spread of the wireless channel. In particular, to effectively combat ISI, the cyclic prefix should be longer than the maximum expected delay spread for the system.

[0058] In an embodiment, cyclic prefixes of different lengths may be used for the OFDM symbols, depending on the expected delay spread. For the specific MIMO WLAN system described above, a cyclic prefix of 400 nsec (8 samples) or 800 nsec (16 samples) may be selected for use for the OFDM symbols. A "short" OFDM symbol uses the 400 nsec cyclic prefix and has a duration of 3.6 μsec. A "long" OFDM symbol uses the 800 nsec cyclic prefix and has a duration of 4.0 μsec. Short OFDM symbols may be used if the maximum expected delay spread is 400 nsec or less, and long OFDM symbols may be used if the delay spread is greater than 400 nsec. Different cyclic prefixes may be selected for use for different transport channels, and the cyclic prefix

may also be dynamically selectable, as described below. Higher system throughput may be achieved by using the shorter cyclic prefix when possible, since more OFDM symbols of shorter duration can be transmitted over a given fixed time interval.

[0059] The MIMO WLAN system may also be designed to not utilize OFDM, and this is within the scope of the invention.

3. Layer Structure

[0060] **FIG. 2** illustrates a layer structure 200 that may be used for the MIMO WLAN system. Layer structure 200 includes (1) applications and upper layer protocols that approximately correspond to Layer 3 and higher of the ISO/OSI reference model (upper layers), (2) protocols and services that correspond to Layer 2 (the link layer), and (3) protocols and services that correspond to Layer 1 (the physical layer).

[0061] The upper layers includes various applications and protocols, such as signaling services 212, data services 214, voice services 216, circuit data applications, and so on. Signaling is typically provided as messages and data is typically provided as packets. The services and applications in the upper layers originate and terminate messages and packets according to the semantics and timing of the communication protocol between the access point and the user terminal. The upper layers utilize the services provided by Layer 2.

[0062] Layer 2 supports the delivery of messages and packets generated by the upper layers. In the embodiment shown in FIG. 2, Layer 2 includes a Link Access Control (LAC) sublayer 220 and a Medium Access Control (MAC) sublayer 230. The LAC sublayer implements a data link protocol that provides for the correct transport and delivery of messages generated by the upper layers. The LAC sublayer utilizes the services provided by the MAC sublayer and Layer 1. The MAC sublayer is responsible for transporting messages and packets using the services provided by Layer 1. The MAC sublayer controls the access to Layer 1 resources by the applications and services in the upper layers. The MAC sublayer may include a Radio Link Protocol (RLP) 232, which is a retransmission mechanism that may be used to provide higher reliability for packet data. Layer 2 provides protocol data units (PDUs) to Layer 1.

[0063] Layer 1 comprises physical layer 240 and supports the transmission and reception of radio signals between the access point and user terminal. The physical layer performs coding, interleaving, modulation, and spatial processing for various transport channels used to send messages and packets generated by the upper layers. In

this embodiment, the physical layer includes a multiplexing sublayer 242 that multiplexes processed PDUs for various transport channels into the proper frame format. Layer 1 provides data in units of frames.

[0064] FIG. 2 shows a specific embodiment of a layer structure that may be used for the MIMO WLAN system. Various other suitable layer structures may also be designed and used for the MIMO WLAN system, and this is within the scope of the invention. The functions performed by each layer are described in further detail below where appropriate.

4. Transport Channels

[0065] A number of services and applications may be supported by the MIMO WLAN system. Moreover, other data required for proper system operation may need to be sent by the access point or exchanged between the access point and user terminals. A number of transport channels may be defined for the MIMO WLAN system to carry various types of data. Table 3 lists an exemplary set of transport channels and also provides a brief description for each transport channel.

Table 3

Transport channels		Description
Broadcast channel	BCH	Used by the access point to transmit pilot and system parameters to the user terminals.
Forward control channel	FCCH	Used by the access point to allocate resources on the downlink and uplink. The resource allocation may be performed on a frame-by-frame basis. Also used to provide acknowledgment for messages received on the RACH.
Forward channel	FCH	Used by the access point to transmit user-specific data to the user terminals and possibly a reference (pilot) used by the user terminals for channel estimation. May also be used in a broadcast mode to send page and broadcast messages to multiple user terminals.
Random access channel	RACH	Used by the user terminals to gain access to the system and send short messages to the access point.
Reverse channel	RCH	Used by the user terminals to transmit data to the access

		point. May also carry a reference used by the access point for channel estimation.

[0066] As shown in Table 3, the downlink transport channels used by the access point includes the BCH, FCCH, and FCH. The uplink transport channels used by the user terminals include the RACH and RCH. Each of these transport channels is described in further detail below.

[0067] The transport channels listed in Table 3 represent a specific embodiment of a channel structure that may be used for the MIMO WLAN system. Fewer, additional, and/or different transport channels may also be defined for use for the MIMO WLAN system. For example, certain functions may be supported by function-specific transport channels (e.g., pilot, paging, power control, and sync channel channels). Thus, other channel structures with different sets of transport channels may be defined and used for the MIMO WLAN system, and this is within the scope of the invention.

5. Frame Structures

[0068] A number of frame structures may be defined for the transport channels. The specific frame structure to use for the MIMO WLAN system is dependent on various factors such as, for example, (1) whether the same or different frequency bands are used for the downlink and uplink and (2) the multiplexing scheme used to multiplex the transport channels together.

[0069] If only one frequency band is available, then the downlink and uplink may be transmitted on different phases of a frame using time division duplexing (TDD), as described below. If two frequency bands are available, then the downlink and uplink may be transmitted on different frequency bands using frequency division duplexing (FDD).

[0070] For both TDD and FDD, the transport channels may be multiplexed together using time division multiplexing (TDM), code division multiplexing (CDM), frequency division multiplexing (FDM), and so on. For TDM, each transport channel is assigned to a different portion of a frame. For CDM, the transport channels are transmitted concurrently but each transport channel is channelized by a different channelization code, similar to that performed in a code division multiple access (CDMA) system. For FDM, each transport channel is assigned a different portion of the frequency band for the link.

[0071] Table 4 lists the various frame structures that may be used to carry the transport channels. Each of these frame structures is described in further detail below. For clarity, the frame structures are described for the set of transport channels listed in Table 3.

Table 4

	Shared frequency band for downlink and uplink	Separate frequency bands for downlink and uplink
Time division	TDD-TDM frame structure	FDD-TDM frame structure
Code division	TDD-CDM frame structure	FDD-CDM frame structure

[0072] **FIG. 3A** illustrates an embodiment of a TDD-TDM frame structure 300a that may be used if a single frequency band is used for both the downlink and uplink. Data transmission occurs in units of TDD frames. Each TDD frame may be defined to span a particular time duration. The frame duration may be selected based on various factors such as, for example, (1) the bandwidth of the operating band, (2) the expected sizes of the PDUs for the transport channels, and so on. In general, a shorter frame duration may provide reduced delays. However, a longer frame duration may be more efficient since header and overhead may represent a smaller fraction of the frame. In a specific embodiment, each TDD frame has a duration of 2 msec.

[0073] Each TDD frame is partitioned into a downlink phase and an uplink phase. The downlink phase is further partitioned into three segments for the three downlink transport channels - the BCH, FCCH, and FCH. The uplink phase is further partitioned into two segments for the two uplink transport channels - the RCH and RACH.

[0074] The segment for each transport channel may be defined to have either a fixed duration or a variable duration that can change from frame to frame. In an embodiment, the BCH segment is defined to have a fixed duration, and the FCCH, FCH, RCH, and RACH segments are defined to have variable durations.

[0075] The segment for each transport channel may be used to carry one or more protocol data units (PDUs) for that transport channel. In the specific embodiment shown in FIG. 3A, a BCH PDU is transmitted in a first segment 310, an FCCH PDU is transmitted in a second segment 320, and one or more FCH PDUs are transmitted in a third segment 330 of the downlink phase. On the uplink phase, one or more RCH PDUs

are transmitted in a fourth segment 340 and one or more RACH PDUs are transmitted in a fifth segment 350 of the TDD frame.

[0076] Frame structure 300a represents a specific arrangement of the various transport channels within a TDD frame. This arrangement can provide certain benefits such as reduced delays for data transmission on the downlink and uplink. The BCH is transmitted first in the TDD frame since it carries system parameters that may be used for the PDUs of the other transport channels within the same TDD frame. The FCCH is transmitted next since it carries channel assignment information indicative of which user terminal(s) are designated to receive downlink data on the FCH and which user terminal(s) are designated to transmit uplink data on the RCH within the current TDD frame. Other TDD-TDM frame structures may also be defined and used for the MIMO WLAN system, and this is within the scope of the invention.

[0077] **FIG. 3B** illustrates an embodiment of an FDD-TDM frame structure 300b that may be used if the downlink and uplink are transmitted using two separate frequency bands. Downlink data is transmitted in a downlink frame 302a, and uplink data is transmitted in an uplink frame 302b. Each downlink and uplink frame may be defined to span a particular time duration (e.g., 2 msec). For simplicity, the downlink and uplink frames may be defined to have the same duration and may further be defined to be aligned at the frame boundaries. However, different frame durations and/or non-aligned (i.e., offset) frame boundaries may also be used for the downlink and uplink.

[0078] As shown in FIG. 3B, the downlink frame is partitioned into three segments for the three downlink transport channels. The uplink frame is partitioned into two segments for the two uplink transport channels. The segment for each transport channel may be defined to have a fixed or variable duration, and may be used to carry one or more PDUs for that transport channel.

[0079] In the specific embodiment shown in FIG. 3B, the downlink frame carries a BCH PDU, an FCCH PDU, and one or more FCH PDUs in segments 310, 320, and 330, respectively. The uplink frame carries one or more RCH PDUs and one or more RACH PDUs in segments 340 and 350, respectively. This specific arrangement may provide the benefits described above (e.g., reduced delays for data transmission). The transport channels may have different PDU formats, as described below. Other FDD-TDM frame structures may also be defined and used for the MIMO WLAN system, and this is within the scope of the invention.

[0080] **FIG. 3C** illustrates an embodiment of an FDD-CDM/FDM frame structure 300c that may also be used if the downlink and uplink are transmitted using separate frequency bands. Downlink data may be transmitted in a downlink frame 304a, and uplink data may be transmitted in an uplink frame 304b. The downlink and uplink frames may be defined to have the same duration (e.g., 2 msec) and aligned at the frame boundaries.

[0081] As shown in FIG. 3C, the three downlink transport channels are transmitted concurrently in the downlink frame, and the two uplink transport channels are transmitted concurrently in the uplink frame. For CDM, the transport channels for each link are "channelized" with different channelization codes, which may be Walsh codes, orthogonal variable spreading factor (OVSF) codes, quasi-orthogonal functions (QOF), and so on. For FDM, the transport channels for each link are assigned different portions of the frequency band for the link. Different amounts of transmit power may also be used for different transport channels in each link.

[0082] Other frame structures may also be defined for the downlink and uplink transport channels, and this is within the scope of the invention. Moreover, it is possible to use different types of frame structure for the downlink and uplink. For example, a TDM-based frame structure may be used for the downlink and a CDM-based frame structure may be used for the uplink.

[0083] In the following description, the MIMO WLAN system is assumed to use one frequency band for both downlink and uplink transmissions. For clarity, the TDD-TDM frame structure shown in FIG. 3A is used for the MIMO WLAN system. For clarity, a specific implementation of the TDD-TDM frame structure is described throughout the specification. For this implementation, the duration of each TDD frame is fixed at 2 msec, and the number of OFDM symbols per TDD frame is a function of the length of the cyclic prefix used for the OFDM symbols. The BCH has a fixed duration of 80 μsec and uses the 800 nsec cyclic prefix for the OFDM symbols transmitted. The remainder of the TDD frame contains 480 OFDM symbols if the 800 nsec cyclic prefix is used, and 533 OFDM symbols plus 1.2 μsec of excess time if the 400 nsec cyclic prefix is used. The excess time can be added to the guard interval at the end of the RACH segment. Other frame structures and other implementations may also be used, and this is within the scope of the invention.

II. Transport Channels

[0084] The transport channels are used to send various types of data and may be categorized into two groups: common transport channels and dedicated transport channels. Because the common and dedicated transport channels are used for different purposes, different processing may be used for these two groups of transport channels, as described in further detail below.

[0085] **Common Transport Channels.** The common transport channels include the BCH, FCCH, and RACH. These transport channels are used to send data to or receive data from multiple user terminals. For improved reliability, the BCH and FCCH are transmitted by the access point using the diversity mode. On the uplink, the RACH is transmitted by the user terminals using the beam-steering mode (if supported by the user terminal). The BCH is operated at a known fixed rate so that the user terminals can receive and process the BCH without any additional information. The FCCH and RACH support multiple rates to allow for greater efficiency. As used herein, each "rate" or "rate set" is associated with a particular code rate (or coding scheme) and a particular modulation scheme.

[0086] **Dedicated Transport Channels.** The dedicated transport channels include the FCH and RCH. These transport channels are normally used to send user-specific data to or by specific user terminals. The FCH and RCH may be dynamically allocated to the user terminals as necessary and as available. The FCH may also be used in a broadcast mode to send overhead, page, and broadcast messages to the user terminals. In general, the overhead, page, and broadcast messages are transmitted prior to any user-specific data on the FCH.

[0087] **FIG. 4** illustrates an exemplary transmission on the BCH, FCCH, FCH, RCH, and RACH based on TDD-TDM frame structure 300a. In this embodiment, one BCH PDU 410 and one FCCH PDU 420 are transmitted in BCH segment 310 and FCCH segment 320, respectively. FCH segment 330 may be used to send one or more FCH PDUs 430, each of which may be intended for a specific user terminal or multiple user terminals. Similarly, one or more RCH PDUs 440 may be sent by one or more user terminals in RCH segment 340. The start of each FCH/RCH PDU is indicated by an FCH/RCH offset from the end of the preceding segment. A number of RACH PDUs 450 may be sent in RACH segment 350 by a number of user terminals to access the system and/or to send short messages, as described below.

[0088] For clarity, the transport channels are described for the specific TDD-TDM frame structure shown in FIGS. 3A and 4.

1. Broadcast Channel (BCH) - Downlink

[0089] The BCH is used by the access point to transmit a beacon pilot, a MIMO pilot, and system parameters to the user terminals. The beacon pilot is used by the user terminals to acquire system timing and frequency. The MIMO pilot is used by the user terminals to estimate the MIMO channel formed by the access point antennas and their own antennas. The beacon and MIMO pilots are described in further detail below. The system parameters specify various attributes of the downlink and uplink transmissions. For example, since the durations of the FCCH, FCH, RACH, and RCH segments are variable, the system parameters that specify the length of each of these segments for the current TDD frame are sent in the BCH.

[0090] **FIG. 5A** illustrates an embodiment of BCH PDU 410. In this embodiment, BCH PDU 410 includes a preamble portion 510 and a message portion 516. Preamble portion 510 further includes a beacon pilot portion 512 and a MIMO pilot portion 514. Portion 512 carries a beacon pilot and has a fixed duration of $T_{CP} = 8\mu\text{sec}$. Portion 514 carries a MIMO pilot and has a fixed duration of $T_{MP} = 32\mu\text{sec}$. Portion 516 carries a BCH message and has a fixed duration of $T_{BM} = 40\mu\text{sec}$. The duration of the BCH PDU is fixed at $T_{CP} + T_{MP} + T_{BM} = 80\mu\text{sec}$.

[0091] A preamble may be used to send one or more types of pilot and/or other information. A beacon pilot comprises a specific set of modulation symbols that is transmitted from all transmit antennas. A MIMO pilot comprises a specific set of modulation symbols that is transmitted from all transmit antennas with different orthogonal codes, which then allows the receivers to recover the pilot transmitted from each antenna. Different sets of modulation symbols may be used for the beacon and MIMO pilots. The generation of the beacon and MIMO pilots is described in further detail below.

[0092] The BCH message carries system configuration information. Table 5 lists the various fields for an exemplary BCH message format.

Table 5 - BCH Message

Fields/ Parameter Names	Length (bits)	Description

Frame Counter	4	TDD frame counter
Net ID	10	Network identifier (ID)
AP ID	6	Access point ID
AP Tx Lvl	4	Access point transmit level
AP Rx Lvl	3	Access point receive level
FCCH Length	6	Duration of FCCH (in units of OFDM symbols)
FCCH Rate	2	Physical layer rate of FCCH
FCH Length	9	Duration of FCH (in units of OFDM symbols)
RCH Length	9	Duration of RCH (in units of OFDM symbols)
RACH Length	5	Duration of RACH (in units of RACH slots)
RACH Slot Size	2	Duration of each RACH slot (in units of OFDM symbols)
RACH Guard Interval	2	Guard interval at the end of RACH
Cyclic Prefix Duration	1	Cyclic prefix duration
Page Bit	1	"0" = page message sent on FCH "1" = no page message sent
Broadcast Bit	1	"0" = broadcast message sent on FCH "1" = no broadcast message sent
RACH Acknowledgment Bit	1	"0" = RACH acknowledgment sent on FCH "1" = no RACH acknowledgment sent
CRC	16	CRC value for the BCH message
Tail Bits	6	Tail bits for convolutional encoder
Reserved	32	Reserved for future use

[0093] The Frame Counter value may be used to synchronize various processes at the access point and user terminals (e.g., the pilot, scrambling codes, cover code, and so on). A frame counter may be implemented with a 4-bit counter that wraps around. This counter is incremented at the start of each TDD frame, and the counter value is included in the Frame Counter field. The Net ID field indicates the identifier (ID) of the network to which the access point belongs. The AP ID field indicates the ID of the access point within the network ID. The AP Tx Lvl and AP Rx Lvl fields indicate the maximum transmit power level and the desired receive power level at the access point,

respectively. The desired receive power level may be used by the user terminal to determine the initial uplink transmit power.

[0094] The FCCH Length, FCH Length, and RCH Length fields indicate the lengths of the FCCH, FCH, and RCH segments, respectively, for the current TDD frame. The lengths of these segments are given in units of OFDM symbols. The OFDM symbol duration for the BCH is fixed at 4.0 μsec. The OFDM symbol duration for all other transport channels (i.e., the FCCH, FCH, RACH, and RCH) is variable and depends on the selected cyclic prefix, which is specified by the Cyclic Prefix Duration field. The FCCH Rate field indicates the rate used for the FCCH for the current TDD frame.

[0095] The RACH Length field indicates the length of the RACH segment, which is given in units of RACH slots. The duration of each RACH slot is given by the RACH Slot Size field, in units of OFDM symbols. The RACH Guard Interval field indicates the amount of time between the last RACH slot and the start of the BCH segment for the next TDD frame. These various fields for the RACH are described in further detail below.

[0096] The Page Bit and Broadcast Bit indicate whether or not page messages and broadcast messages, respectively, are being sent on the FCH in the current TDD frame. These two bits may be set independently for each TDD frame. The RACH Acknowledgment Bit indicates whether or not acknowledgments for PDUs sent on the RACH in prior TDD frames are being sent on the FCCH in the current TDD frame.

[0097] The CRC field includes a CRC value for the entire BCH message. This CRC value may be used by the user terminals to determine whether the received BCH message is decoded correctly (i.e., good) or in error (i.e., erased). The Tail Bits field includes a group of zeros used to reset the convolutional encoder to a known state at the end of the BCH message.

[0098] As shown in Table 5, the BCH message includes a total of 120 bits. These 120 bits may be transmitted with 10 OFDM symbols using the processing described in detail below.

[0099] Table 5 shows a specific embodiment of the format for the BCH message. Other BCH message formats with fewer, additional, and/or different fields may also be defined and used, and this is within the scope of the invention.

2. Forward Control Channel (FCCH) - Downlink

[00100] In an embodiment, the access point is able to allocate resources for the FCH and RCH on a per frame basis. The FCCH is used by the access point to convey the resource allocation for the FCH and RCH (i.e., the channel assignments).

[00101] **FIG. 5B** illustrates an embodiment of FCCH PDU 420. In this embodiment, the FCCH PDU includes only a portion 520 for an FCCH message. The FCCH message has a variable duration that can change from frame to frame, depending on the amount of scheduling information being carried on the FCCH for that frame. The FCCH message duration is in even number of OFDM symbols and given by the FCCH Length field on the BCH message. The duration of messages sent using the diversity mode (e.g., BCH and FCCH messages) is given in even number of OFDM symbols because the diversity mode transmits OFDM symbols in pairs, as described below.

[00102] In an embodiment, the FCCH can be transmitted using four possible rates. The specific rate used for the FCCH PDU in each TDD frame is indicated by the FCCH Phy Mode field in the BCH message. Each FCCH rate corresponds to a particular code rate and a particular modulation scheme and is further associated with a particular transmission mode, as shown in Table 26.

[00103] An FCCH message may include zero, one, or multiple information elements (IEs). Each information element may be associated with a specific user terminal and may be used to provide information indicative of the assignment of FCH/RCH resources for that user terminal. Table 6 lists the various fields for an exemplary FCCH message format.

Table 6 - FCCH Message

Fields/ Parameter Names	**Length** (bits)	**Description**
N_IE	6	Number of IEs included in the FCCH message
N_IE information elements, each including:		
IE Type	4	IE type
MAC ID	10	ID assigned to the user terminal
Control Fields	48 or 72	Control fields for channel assignment
Padding Bits	Variable	Pad bits to achieve even number of OFDM symbols in the FCCH message

CRC	16	CRC value for the FCCH message
Tail Bits	6	Tail bits for convolutional encoder

[00104] The N_IE field indicates the number of information elements included in the FCCH message sent in the current TDD frame. For each information element (IE) included in the FCCH message, the IE Type field indicates the particular type of this IE. A number of IE types are defined for use to allocate resources for different types of transmissions, as described below.

[00105] The MAC ID field identifies the specific user terminal for which the information element is intended. Each user terminal registers with the access point at the start of a communication session and is assigned a unique MAC ID by the access point. This MAC ID is used to identify the user terminal during the session.

[00106] The Control Fields are used to convey channel assignment information for the user terminal and are described in detail below. The Padding Bits field includes a sufficient number of padding bits so that the overall length of the FCCH message is an even number of OFDM symbols. The FCCH CRC field includes a CRC value that may be used by the user terminals to determine whether the received FCCH message is decoded correctly or in error. The Tail Bits field includes zeros used to reset the convolutional encoder to a known state at the end of the FCCH message. Some of these fields are described in further detail below.

[00107] A number of transmission modes are supported by the MIMO WLAN system for the FCH and RCH, as indicated in Table 1. Moreover, a user terminal may be active or idle during a connection. Thus, a number of types of IE are defined for use to allocate FCH/RCH resources for different types of transmissions. Table 7 lists an exemplary set of IE types.

Table 7 - FCCH IE Types

IE Type	IE Size (bits)	IE Type	Description
0	48	Diversity Mode	Diversity mode only
1	72	Spatial Multiplexing Mode	Spatial multiplexing mode - variable rate services
2	48	Idle Mode	Idle state - variable rate services

3	48	RACH Acknowledgment	RACH acknowledgment – diversity mode
4		Beam Steering Mode	Beam steering mode
5-15	-	Reserved	Reserved for future use

[00108] For IE types 0, 1 and 4, resources are allocated to a specific user terminal for both the FCH and RCH (i.e., in channel pairs). For IE type 2, minimal resources are allocated to the user terminal on the FCH and RCH to maintain up-to-date estimate of the link. An exemplary format for each IE type is described below. In general, the rates and durations for the FCH and RCH can be independently assigned to the user terminals.

A. <u>IE Type 0, 4 – Diversity/Beam-Steering Mode</u>

[00109] IE type 0 and 4 are used to allocate FCH/RCH resources for the diversity and beam-steering modes, respectively. For fixed low-rate services (e.g., voice), the rate remains fixed for the duration of the call. For variable rate services, the rate may be selected independently for the FCH and RCH. The FCCH IE indicates the location of the FCH and RCH PDUs assigned to the user terminal. Table 8 lists the various fields of an exemplary IE Type 0 and 4 information element.

Table 8 - FCCH IE Type 0 and 4

Fields/ Parameter Names	**Length** (bits)	**Description**
IE Type	4	IE type
MAC ID	10	Temporary ID assigned to the user terminal
FCH Offset	9	FCH offset from start of the TDD frame (in OFDM symbols)
FCH Preamble Type	2	FCH preamble size (in OFDM symbols)
FCH Rate	4	Rate for the FCH
RCH Offset	9	RCH offset from start of the TDD frame (in OFDM symbols)
RCH Preamble Type	2	RCH preamble size (in OFDM symbols)
RCH Rate	4	Rate for the RCH
RCH Timing Adjustment	2	Timing adjustment parameter for RCH

RCH Power Control	2	Power control bits for RCH

[00110] The FCH and RCH Offset fields indicate the time offset from the beginning of the current TDD frame to the start of the FCH and RCH PDUs, respectively, assigned by the information element. The FCH and RCH Rate fields indicate the rates for the FCH and RCH, respectively.

[00111] The FCH and RCH Preamble Type fields indicate the size of the preamble in the FCH and RCH PDUs, respectively. Table 9 lists the values for the FCH and RCH Preamble Type fields and the associated preamble sizes.

Table 9 - Preamble Type

Type	Bits	Preamble Size
0	00	0 OFDM symbol
1	01	1 OFDM symbol
2	10	4 OFDM symbols
3	11	8 OFDM symbols

[00112] The RCH Timing Adjustment field includes two bits used to adjust the timing of the uplink transmission from the user terminal identified by the MAC ID field. This timing adjustment is used to reduce interference in a TDD-based frame structure (such as the one shown in FIG. 3A) where the downlink and uplink transmissions are time division duplexed. Table 10 lists the values for the RCH Timing Adjustment field and the associated actions.

Table 10 - RCH Timing Adjustment

Bits	Description
00	Maintain current timing
01	Advance uplink transmit timing by 1 sample
10	Delay uplink transmit timing by 1 sample
11	Not used

[00113] The RCH Power Control field includes two bits used to adjust the transmit power of the uplink transmission from the identified user terminal. This power control

is used to reduce interference on the uplink. Table 11 lists the values for the RCH Power Control field and the associated actions.

Table 11 - RCH Power Control

Bits	Description
00	Maintain current transmit power
01	Increase uplink transmit power by δ dB, where δ is a system parameter.
10	Decrease uplink transmit power by δ dB, where δ is a system parameter.
11	Not used

[00114] The channel assignment for the identified user terminal may be provided in various manners. In an embodiment, the user terminal is assigned FCH/RCH resources for only the current TDD frame. In another embodiment, the FCH/RCH resources are assigned to the terminal for each TDD frame until canceled. In yet another embodiment, the FCH/RCH resources are assigned to the user terminal for every n-th TDD frame, which is referred to as "decimated" scheduling of TDD frames. The different types of assignment may be indicated by an Assignment Type field in the FCCH information element.

B. IE Type 1 - Spatial Multiplexing Mode

[00115] IE type 1 is used to allocate FCH/RCH resources to user terminals using the spatial multiplexing mode. The rate for these user terminals is variable, and may be selected independently for the FCH and RCH. Table 12 lists the various fields of an exemplary IE type 1 information element.

Table 12 - FCCH IE Type 1

Fields/ Parameter Names	Length (bits)	Description
IE Type	4	IE type
MAC ID	10	Temporary ID assigned to the user terminal
FCH Offset	9	FCH offset from end of FCCH (in OFDM symbols)
FCH Preamble Type	2	FCH preamble size (in OFDM symbols)

FCH Spatial Channel 1 Rate	4	Rate for the FCH for spatial channel 1
FCH Spatial Channel 2 Rate	4	Rate for the FCH for spatial channel 2
FCH Spatial Channel 3 Rate	4	Rate for the FCH for spatial channel 3
FCH Spatial Channel 4 Rate	4	Rate for the FCH for spatial channel 4
RCH Offset	9	RCH offset from end of FCH (in OFDM symbols)
RCH Preamble Type	2	RCH preamble size (in OFDM symbols)
RCH Spatial Channel 1 Rate	4	Rate for the RCH for spatial channel 1
RCH Spatial Channel 2 Rate	4	Rate for the RCH for spatial channel 2
RCH Spatial Channel 3 Tate	4	Rate for the RCH for spatial channel 3
RCH Spatial Channel 4 Rate	4	Rate for the RCH for spatial channel 4
RCH Timing Adjustment	2	Timing adjustment parameter for RCH
Reserved	2	Reserved for future use

[00116] For IE type 1, the rate for each spatial channel may be selected independently on the FCH and RCH. The interpretation of the rates for the spatial multiplexing mode is general in that it can specify the rate per spatial channel (e.g., for up to four spatial channels for the embodiment shown in Table 12). The rate is given per eigenmode if the transmitter performs spatial processing to transmit data on the eigenmodes. The rate is given per antenna if the transmitter simply transmits data from the transmit antennas and the receiver performs the spatial processing to isolate and recover the data (for the non-steered spatial multiplexing mode).

[00117] The information element includes the rates for all enabled spatial channels and zeros for the ones not enabled. User terminals with less than four transmit antennas set the unused FCH/RCH Spatial Channel Rate fields to zero. Since the access point is equipped with four transmit/receive antennas, user terminals with more than four transmit antennas may use them to transmit up to four independent data streams.

C. IE Type 2 - Idle Mode

[00118] IE type 2 is used to provide control information for user terminals operating in an *Idle* state (described below). In an embodiment, when a user terminal is in the *Idle* state, steering vectors used by the access point and user terminal for spatial processing

are continually updated so that data transmission can start quickly if and when resumed. Table 13 lists the various fields of an exemplary IE type 2 information element.

Table 13 - FCCH IE Type 2

Fields/ Parameter Names	Length (bits)	Description
IE Type	4	IE type
MAC ID	10	Temporary ID assigned to the user terminal
FCH Offset	9	FCH offset from end of FCCH (in OFDM symbols)
FCH Preamble Type	2	FCH preamble size (in OFDM symbols)
RCH Offset	9	RCH offset from end of FCH (in OFDM symbols)
RCH Preamble Type	2	RCH preamble size (in OFDM symbols)
Reserved	12	Reserved for future use

D. IE Type 3 - RACH Quick Acknowledgment

[00119] IE type 3 is used to provide quick acknowledgment for user terminals attempting to access the system via the RACH. To gain access to the system or to send a short message to the access point, a user terminal may transmit an RACH PDU on the uplink. After the user terminal sends the RACH PDU, it monitors the BCH to determine if the RACH Acknowledgement Bit is set. This bit is set by the access point if any user terminal was successful in accessing the system and an acknowledgment is being sent for at least one user terminal on the FCCH. If this bit is set, then the user terminal processes the FCCH for acknowledgment sent on the FCCH. IE Type 3 information elements are sent if the access point desires to acknowledge that it correctly decoded the RACH PDUs from the user terminals without assigning resources. Table 14 lists the various fields of an exemplary IE Type 3 information element.

Table 14 - FCCH IE Type 3

Fields/ Parameter Names	Length (bits)	Description
IE Type	4	IE type
MAC ID	10	Temporary ID assigned to user terminal
Reserved	34	Reserved for future use

[00120] A single or multiple types of acknowledgment may be defined and sent on the FCCH. For example, a quick acknowledgment and an assignment-based acknowledgment may be defined. A quick acknowledgment may be used to simply acknowledge that the RACH PDU has been received by the access point but that no FCH/RCH resources have been assigned to the user terminal. An assignment-based acknowledgment includes assignments for the FCH and/or RCH for the current TDD frame.

[00121] The FCCH may be implemented in other manners and may also be transmitted in various ways. In one embodiment, the FCCH is transmitted at a single rate that is signaled in the BCH message. This rate may be selected, for example, based on the lowest signal-to-noise-and-interference ratios (SNRs) of all user terminals for which the FCCH is being sent in the current TDD frame. Different rates may be used for different TDD frames depending on the channel conditions of the recipient user terminals in each TDD frame.

[00122] In another embodiment, the FCCH is implemented with multiple (e.g., four) FCCH subchannels. Each FCCH subchannel is transmitted at a different rate and is associated with a different required SNR in order to recover the subchannel. The FCCH subchannels are transmitted in order from lowest rate to highest rate. Each FCCH subchannel may or may not be transmitted in a given TDD frame. The first FCCH subchannel (with the lowest rate) is transmitted first and can be received by all user terminals. This FCCH subchannel can indicate whether or not each of the remaining FCCH subchannels will be transmitted in the current TDD frame. Each user terminal can process the transmitted FCCH subchannels to obtain its FCCH information element. Each user terminal can terminate processing of the FCCH if any of the following occurs: (1) failure to decode the current FCCH subchannel, (2) reception of its FCCH information element in the current FCCH subchannel, or (3) all transmitted FCCH subchannels have been processed. A user terminal can terminate processing of the FCCH as soon as it encounters FCCH decoding failure because the FCCH subchannels are transmitted at ascending rates and the user terminal is unlikely to be able to decode subsequent FCCH subchannels transmitted at higher rates.

3. Random Access Channel (RACH) - Uplink

[00123] The RACH is used by the user terminals to gain access to the system and to send short messages to the access point. The operation of the RACH is based on a slotted Aloha random access protocol, which is described below.

[00124] **FIG. 5C** illustrates an embodiment of RACH PDU 450. In this embodiment, the RACH PDU includes a preamble portion 552 and a message portion 554. Preamble portion 552 may be used to send a steered reference (i.e., a steered pilot), if the user terminal is equipped with multiple antennas. The steered reference is a pilot comprised of a specific set of modulation symbols that is subjected to spatial processing prior to transmission on the uplink. The spatial processing allows the pilot to be transmitted on a specific eigenmode of the MIMO channel. The processing for the steered reference is described in further detail below. Preamble portion 552 has a fixed duration of at least 2 OFDM symbols. Message portion 554 carries an RACH message and has a variable duration. The duration of the RACH PDU is thus variable.

[00125] In an embodiment, four different rates are supported for the RACH. The specific rate used for each RACH message is indicated by a 2-bit RACH data rate indicator (DRI), which is embedded in the preamble portion of the RACH PDU, as shown in FIG. 5C. In an embodiment, four different message sizes are also supported for the RACH. The size of each RACH message is indicated by a Message Duration field included in the RACH message. Each RACH rate supports 1, 2, 3 or all 4 message sizes. Table 15 lists the four RACH rates, their associated coding and modulation parameters, and the message sizes supported by these RACH rates.

Table 15

RACH Rates				**RACH Message Sizes** (in bits and OFDM symbols)			
bps/Hz	**Code Rate**	**Modulation**	**DRI**	96 bits	192 bits	384 bits	768 bits
0.25	0.25	BPSK	(1,1)	8	n/a	n/a	n/a
0.5	0.5	BPSK	(1,-1)	4	8	n/a	n/a
1	0.5	QPSK	(-1, 1)	2	4	8	n/a
2	0.5	16 QAM	(-1, -1)	1	2	4	8

[00126] The RACH message carries short messages and access requests from the user terminal. Table 16 lists the various fields of an exemplary RACH message format and the size of each field for each of the four different message sizes.

Table 16

Fields/ Parameter Names	RACH Message Sizes				Description
	96 bits	192 bits	384 bits	768 bits	
Message Duration	2	2	2	2	Duration of message
MAC PDU Type	4	4	4	4	RACH message type
MAC ID	10	10	10	10	MAC ID
Slot ID	6	6	6	6	Tx Slot ID
Payload	44	140	332	716	Info bits
CRC	24	24	24	24	CRC value for the RACH message
Tail Bits	6	6	6	6	Tail bits

[00127] The Message Duration field indicates the size of the RACH message. The MAC PDU Type field indicates the RACH message type. The MAC ID field contains the MAC ID that uniquely identifies the user terminal sending the RACH message. During initial system access, a unique MAC ID has not been assigned to the user terminal. In this case, a registration MAC ID (e.g., a specific value reserved for registration purpose) may be included in the MAC ID field. The Slot ID field indicates the starting RACH slot on which the RACH PDU was sent (the RACH timing and transmission is described below). The Payload field includes the information bits for the RACH message. The CRC field contains a CRC value for the RACH message, and the Tail Bits field is used to reset the convolutional encoder for the RACH. The operation of the RACH in conjunction with the BCH and FCCH for system access is described in further detail below.

[00128] The RACH may also be implemented with a "fast" RACH (F-RACH) and a "slow" RACH (S-RACH). The F-RACH and S-RACH can be designed to efficiently support user terminals in different operating states. For example, the F-RACH may be used by user terminals that (1) have registered with the system, (2) can compensate for their round trip delays (RTDs) by properly advancing their transmit timing, and (3)

achieve the required SNR for operation on the F-RACH. The S-RACH may be used by user terminals that cannot use the F-RACH for any reasons.

[00129] Different designs may be used for the F-RACH and S-RACH to facilitate rapid access to the system whenever possible and to minimize the amount of system resources needed to implement random access. For example, the F-RACH can use a shorter PDU, employ a weaker coding scheme, require F-RACH PDUs to arrive approximately time-aligned at the access point, and utilize a slotted Aloha random access scheme. The S-RACH can use a longer PDU, employ a stronger coding scheme, allow S-RACH PDUs to arrive non-aligned in time at the access point, and utilize an unslotted Aloha random access scheme.

[00130] For simplicity, the following description assumes that a single RACH is used for the MIMO WLAN system.

4. Forward Channel (FCH) - Downlink

[00131] The FCH is used by the access point to transmit user-specific data to specific user terminals and page/broadcast messages to multiple user terminals. The FCH may also be used to transmit pilot to user terminals. The FCH can be allocated on a per frame basis. A number of FCH PDU types are provided to accommodate different uses of the FCH. Table 17 lists an exemplary set of FCH PDU types.

Table 17 - FCH PDU Types

Code	FCH PDU Type	Description
0	Message Only	FCH broadcast/page service/user message
1	Message and Preamble	FCH user message
2	Preamble Only	FCH *Idle* state

[00132] FCH PDU Type 0 is used to send page/broadcast messages and user messages/packets on the FCH and only includes the message/packet. (Data for a specific user terminal may be sent as a message or a packet, and these two terms are used interchangeably herein.) FCH PDU Type 1 is used to send user packets and includes a preamble. FCH PDU Type 2 includes only the preamble and no message/packet, and is associated with *Idle* state FCH traffic.

[00133] **FIG. 5D** illustrates an embodiment of an FCH PDU 430a for FCH PDU Type 0. In this embodiment, FCH PDU 430a includes only a message portion 534a for a page/broadcast message or a user packet. The message/packet can have variable length,

which is given by the FCH Message Length field in the FCH PDU. The message length is given in integer number of PHY frames (described below). The rate and transmission mode for the page/broadcast message are specified and described below. The rate and transmission mode for the user packet are specified in the associated FCCH information element.

[00134] **FIG. 5E** illustrates an embodiment of an FCH PDU 430b for FCH PDU Type 1. In this embodiment, FCH PDU 430b includes a preamble portion 532b and a message/packet portion 534b. Preamble portion 532b is used to send a MIMO pilot or a steered reference and has a variable length, which is given by the FCH Preamble Type field in the associated FCCH information element. Portion 534b is used to send an FCH packet and also has a variable length (in integer number of PHY frames), which is given by the FCH Message Length field in the FCH PDU. The FCH packet is sent using the rate and transmission mode specified by the associated FCCH information element.

[00135] **FIG. 5F** illustrates an embodiment of an FCH PDU 430c for FCH PDU Type 2. In this embodiment, FCH PDU 430c includes only a preamble portion 532c and no message portion. The length of the preamble portion is indicated by the FCCH IE. The FCH PDU Type 2 may be used to allow the user terminal to update its channel estimate while in the *Idle* state.

[00136] A number of FCH Message types are provided to accommodate different uses of the FCH. Table 18 lists an exemplary set of FCH Message types.

Table 18 - FCH Message Types

Code	FCH Message Type	Description
0	Page Message	Page message - diversity mode, rate = 0.25 bps/Hz
1	Broadcast Message	Broadcast message - diversity mode, rate = 0.25 bps/Hz
2	User Packet	Dedicated channel operation - user terminal specific PDU, rate specified in the FCCH
3-15	Reserved	Reserved for future use

[00137] A page message may be used to page multiple user terminals and is sent using FCH PDU Type 0. If the Page Bit in the BCH message is set, then one or more FCH PDUs with page messages (or "Page PDUs") are sent first on the FCH. Multiple Page PDUs may be sent in the same frame. Page PDUs are transmitted using the diversity

mode and the lowest rate of 0.25 bps/Hz to increase the likelihood of correct reception by the user terminals.

[00138] A broadcast message may be used to send information to multiple user terminals and is sent using FCH PDU Type 0. If the Broadcast Bit in the BCH message is set, then one or more FCH PDUs with broadcast messages (or "Broadcast PDUs") are sent on the FCH immediately following any Page PDUs sent on the FCH. The Broadcast PDUs are also transmitted using the diversity mode and the lowest rate of 0.25 bps/Hz to increase the likelihood of correct reception.

[00139] A user packet may be used to send user-specific data, and may be sent using FCH PDU Type 1 or 2. User PDUs of Type 1 and 2 are sent on the FCH following any Page and Broadcast PDUs sent on the FCH. Each User PDU may be transmitted using the diversity, beam-steering, or spatial multiplexing mode. The FCCH information element specifies the rate and transmission mode used for each User PDU sent on the FCH.

[00140] A message or packet sent on the FCH comprises an integer number of PHY frames. In an embodiment and as described below, each PHY frame may include a CRC value that permits individual PHY frames in an FCH PDU to be checked and retransmitted if necessary. For asynchronous services, the RLP may be employed for segmentation, retransmission, and reassembly of PHY frames within a given FCH PDU. In another embodiment, a CRC value is provided for each message or packet, instead of each PHY frame.

[00141] **FIG. 6** illustrates an embodiment of the structure for an FCH packet 534. The FCH packet comprises an integer number of PHY frames 610. Each PHY frame includes a payload field 622, a CRC field 624, and a tail bit field 626. The first PHY frame for the FCH packet further includes a header field 620, which indicates the message type and duration. The last PHY frame in the FCH packet further includes a pad bit field 628, which contains zero padding bits at the end of the payload in order to fill the last PHY frame. In an embodiment, each PHY frame comprises 6 OFDM symbols. The number of bits included in each PHY frame is dependent on the rate used for that PHY frame.

[00142] Table 19 lists the various fields for an exemplary FCH PDU format for FCH PDU Types 0 and 1.

Table 19 - FCH PDU Format

	Fields/ Parameter Names	Length (bits)	Description
First PHY frame	FCH Message Type	4	FCH message type
	FCH Message Length	16	Number of bytes in FCH PDU
	Payload	Variable	Payload bits
	CRC	16	CRC value for PHY frame (optional)
	Tail Bits	6	Tail bits for convolutional encoder
Each Middle PHY frame	Payload	Variable	Payload bits
	CRC	16	CRC value for PHY frame (optional)
	Tail Bits	6	Tail bits for convolutional encoder
Last PHY frame	Payload	Variable	Payload bits
	Pad bits	Variable	Pad bits to fill out PHY frame
	CRC	16	CRC value for PHY frame (optional)
	Tail Bits	6	Tail bits for convolutional encoder

The FCH Message Type and FCH Message Length fields are sent in the header of the first PHY frame of the FCH PDU. The payload, CRC, and tail bits fields are included in each PHY frame. The payload portion of each FCH PDU carries the information bits for the page/broadcast message or user-specific packet. Pad bits are used to fill the last PHY frame of the FCH PDU, if required.

[00143] A PHY frame may also be defined to comprise some other number of OFDM symbols (e.g., one, two, four, eight, and so on). The PHY frame may be defined with even number of OFDM symbols because OFDM symbols are transmitted in pairs for the diversity mode, which may be used for the FCH and RCH. The PHY frame size may be selected based on the expected traffic such that inefficiency is minimized. In particular, if the frame size is too large, then inefficiency results from using a large PHY frame to send a small amount of data. Alternatively, if the frame size is too small, then the overhead represents a larger fraction of the frame.

5. <u>Reverse Channel (RCH) - Uplink</u>

[00144] The RCH is used by the user terminals to transmit uplink data and pilot to the access point. The RCH may be allocated on a per TDD frame basis. One or more user terminals may be designated to transmit on the RCH in any given TDD frame. A

number of RCH PDU types are provided to accommodate different operating modes on the RCH. Table 20 lists an exemplary set of RCH PDU types.

Table 20 - RCH PDU Types

Code	RCH PDU Type	Description
0	Message Only	RCH user message, no preamble
1	Message and Preamble, not Idle	RCH user message, with preamble
2	Message and Preamble, Idle	RCH Idle state message with preamble

[00145] RCH PDU Type 0 is used to send a message/packet on the RCH and does not include a preamble. RCH PDU Type 1 is used to send a message/packet and includes a preamble. RCH PDU Type 2 includes a preamble and a short message, and is associated with *Idle* state RCH traffic.

[00146] **FIG. 5D** illustrates an embodiment of an RCH PDU for RCH PDU Type 0. In this embodiment, the RCH PDU includes only a message portion 534a for a variable-length RCH packet, which is given in integer number of PHY frames by the RCH Message Length field in the RCH PDU. The rate and transmission mode for the RCH packet are specified in the associated FCCH information element.

[00147] **FIG. 5E** illustrates an embodiment of an RCH PDU for RCH PDU Type 1. In this embodiment, the RCH PDU includes a preamble portion 532b and a packet portion 534b. Preamble portion 532b is used to send a reference (e.g., a MIMO pilot or a steered reference) and has a variable length, which is given by the RCH Preamble Type field in the associated FCCH information element. Portion 534b is used to send an RCH packet and also has a variable length, which is given by the RCH Message Length field in the RCH PDU. The RCH packet is sent using the rate and transmission mode specified in the associated FCCH information element.

[00148] **FIG. 5G** illustrates an embodiment of an RCH PDU 350d for RCH PDU Type 2. In this embodiment, the RCH PDU includes a preamble portion 532d and a message portion 536d. Preamble portion 532d is used to send a reference and has a length of 1, 4 or 8 OFDM symbols. Portion 536d is used to send a short RCH message and has a fixed length of one OFDM symbol. The short RCH message is sent using a specific rate and transmission mode (e.g., rate 1/2 or rate 1/4 and BPSK modulation).

[00149] A packet sent on the RCH (for PDU Types 0 and 1) comprises an integer number of PHY frames. The structure for an RCH packet (for PDU Types 0 and 1) is

shown in FIG. 6, and is the same as for the FCH packet. The RCH packet comprises an integer number of PHY frames 610. Each PHY frame includes payload field 622, an optional CRC field 624, and tail bit field 626. The first PHY frame in the RCH packet further includes header field 620, and the last PHY frame in the packet further includes pad bit field 628.

[00150] Table 21 lists the various fields for an exemplary RCH PDU format for RCH PDU Types 0 and 1.

Table 21 - RCH PDU Format (PDU Types 0 and 1)

	Fields/ Parameter Names	Length (bits)	Description
First PHY frame	RCH Message Type	4	RCH message type
	RCH Message Length	16	Number of bytes in RCH PDU
	FCH Rate Indicator	16	Indicate maximum rate for each spatial channel on FCH
	Payload	Variable	Payload bits
	CRC	16	CRC value for PHY frame (optional)
	Tail Bits	6	Tail bits for convolutional encoder
Each Middle PHY frame	Payload	Variable	Payload bits
	CRC	16	CRC value for PHY frame (optional)
	Tail Bits	6	Tail bits for convolutional encoder
Last PHY frame	Payload	Variable	Payload bits
	Pad bits	Variable	Pad bits to fill out PHY frame
	CRC	16	CRC value for PHY frame (optional)
	Tail Bits	6	Tail bits for convolutional encoder

The RCH Message Type, RCH Message Length, and FCH Rate Indicator fields are sent in the header of the first PHY frame of the RCH PDU. The FCH Rate Indicator field is used to convey FCH rate information (e.g., the maximum rates supported by each of the spatial channels) to the access point.

[00151] Table 22 lists the various fields for an exemplary RCH PDU format for RCH PDU Type 2.

Table 22 - RCH Message for RCH PDU Type 2

Fields/ Parameter Names	Length (bits)	Description
FCH Rate Indicator	16	Indicate maximum rate for each spatial channel on FCH
RCH Request	1	User terminal request to send additional data
Reserved	1	Reserved for future use
Tail Bits	6	Tail bits for convolutional encoder

The RCH Request field is used by the user terminal to request additional capacity on the uplink. This short RCH message does not include a CRC and is transmitted in a single OFDM symbol.

6. Dedicated Channel Activity

[00152] Data transmission on the FCH and RCH can occur independently. Depending on the transmission modes selected for use for the FCH and RCH, one or multiple spatial channels (for the beam-steering and diversity modes) may be active and used for data transmission for each dedicated transport channel. Each spatial channel may be associated with a specific rate.

[00153] When only the FCH or only the RCH has all four rates set to zero, the user terminal is idle on that link. The idle terminal still transmits an idle PDU on the RCH. When both the FCH and RCH have all four rates set to zero, both the access point and user terminal are off and not transmitting. User terminals with less than four transmit antennas set the unused rate fields to zero. User terminals with more than four transmit antennas use no more than four spatial channels to transmit data. Table 23 shows the transmission state and channel activity when the rates on all four spatial channels of either the FCH or RCH (or both) are set to zero.

Table 23

FCH Rates	RCH Rates	Channel Activity	Transmission State
At least one rate on FCH $\neq 0$	At least one rate on RCH $\neq 0$	FCH and RCH are active	FCH and/or RCH are transmitting
At least one rate on FCH $\neq 0$	All rates on RCH = 0	FCH active, RCH idle	

All rates on FCH = 0	At least one rate on RCH ≠ 0	FCH idle, RCH active	
All rates on FCH = 0	All rates on RCH = 0	FCH and RCH are OFF	No transmissions

[00154] There may also be a situation where both the RCH and FCH are idle (i.e., not transmitting data) but still transmitting preamble. This is referred to as the *Idle* state. The control fields used to support a user terminal in the *Idle* state are provided in an FCCH IE Type 2 information element, which is shown in Table 13.

7. Alternative Designs

[00155] For clarity, specific PDU types, PDU structures, message formats, and so on, have been described for an exemplary design. Fewer, additional, and/or different types, structures, and formats may also be defined for use, and this is within the scope of the invention.

III. OFDM Subband Structures

[00156] In the above description, the same OFDM subband structure is used for all of the transport channels. Improved efficiency may be achieved by using different OFDM subband structures for different transport channels. For example, a 64-subband structure may be used for some transport channels, a 256-subband structure may be used for some other transport channels, and so on. Moreover, multiple OFDM subband structures may be used for a given transport channel.

[00157] For a given system bandwidth of W, the duration of an OFDM symbol is dependent on the number of total subbands. If the total number of subbands is N, then the duration of each transformed symbol (without a cyclic prefix) is N/W μsec (if W is given in MHz). A cyclic prefix is added to each transformed symbol to form a corresponding OFDM symbol. The length of the cyclic prefix is determined by the expected delay spread of the system. The cyclic prefix represents overhead, which is needed for each OFDM symbol in order to combat a frequency selective channel. This overhead represents a larger percentage of the OFDM symbol if the symbol is short and a smaller percentage if the symbol is long.

[00158] Since different transport channels may be associated with different types of traffic data, a suitable OFDM subband structure may be selected for use for each

transport channel to match the expected traffic data type. If a large amount of data is expected to be transmitted on a given transport channel, then a larger subband structure may be defined for use for the transport channel. In this case, the cyclic prefix would represent a smaller percentage of the OFDM symbol and greater efficiency may be achieved. Conversely, if a small amount of data is expected to be transmitted on a given transport channel, than a smaller subband structure may be defined for use for the transport channel. In this case, even though the cyclic prefix represents a larger percentage of the OFDM symbol, greater efficiency may still be achieved by reducing the amount of excess capacity by using a smaller size OFDM symbol. The OFDM symbol may thus be viewed as a "box car" that is used to send data, and the proper size "box car" may be selected for each transport channel depending on the amount of data expected to be sent.

[00159] For example, for the embodiment described above, the data on the FCH and RCH is sent in PHY frames, each of which comprises 6 OFDM symbols. In this case, another OFDM structure may be defined for use for the FCH and RCH. For example, a 256-subband structure may be defined for the FCH and RCH. A "large" OFDM symbol for the 256-subband structure would be approximately four times the duration of a "small" OFDM symbol for the 64-subband structure but would have four times the data-carrying capacity. However, only one cyclic prefix is needed for one large OFDM symbol, whereas four cyclic prefixes are needed for the equivalent four small OFDM symbols. Thus, the amount of overhead for the cyclic prefixes may be reduced by 75% by the use of the larger 256-subband structure.

[00160] This concept may be extended so that different OFDM subband structures may be used for the same transport channel. For example, the RCH supports different PDU types, each of which may be associated with a certain size. In this case, a larger subband structure may be used for a larger-size RCH PDU type, and a smaller subband structure may be used for a smaller-size RCH PDU type. A combination of different subband structures may also be used for a given PDU. For example, if one long OFDM symbol is equivalent to four short OFDM symbols, then a PDU may be sent using N_{large} large OFDM symbols and N_{small} small OFDM symbols, where $N_{large} \geq 0$ and $3 \geq N_{small} \geq 0$.

[00161] Different OFDM subband structures are associated with OFDM symbols of different lengths. Thus, if different OFDM subband structures are used for different transport channels (and/or for the same transport channel), then the FCH and RCH offsets for the FCH and RCH PDUs would need to be specified with the proper time resolution, which is smaller than an OFDM symbol period. In particular, the time increment for the FCH and RCH PDUs may be given in integer numbers of cyclic prefix length, instead of OFDM symbol period.

IV. Rates and Transmission Modes

[00162] The transport channels described above are used to send various types of data for various services and functions. Each transport channel may be designed to support one or more rates and one or more transmission modes.

1. Transmission Modes

[00163] A number of transmission modes are supported for the transport channels. Each transmission mode is associated with specific spatial processing at the transmitter and receiver, as described below. Table 24 lists the transmission mode(s) supported by each of the transport channels.

Table 24

Transport Channels	Transmission Modes			
	SIMO	Tx Diversity	Beam-Steering	Spatial Multiplexing
BCH	-	X		-
FCCH	-	X	-	-
RACH	X	-	X	-
FCH	-	X	X	X
RCH	X	X	X	X

For the diversity mode, each data symbol is transmitted redundantly over multiple transmit antennas, multiple subbands, multiple symbol periods, or a combination thereof to achieve spatial, frequency, and/or time diversity. For the beam-steering mode, a single spatial channel is used for data transmission (typically the best spatial channel), and each data symbol is transmitted on the single spatial channel using full transmit power available for the transmit antennas. For the spatial multiplexing mode, multiple

spatial channels are used for data transmission, and each data symbol is transmitted on one spatial channel, where a spatial channel may correspond to an eigenmode, a transmit antenna, and so on. The beam-steering mode may be viewed as a special case of the spatial multiplexing mode whereby only one spatial channel is used for data transmission.

[00164] The diversity mode may be used for the common transport channels (BCH and FCCH) for the downlink from the access point to the user terminals. The diversity mode may also be used for the dedicated transport channels (FCH and RCH). The use of the diversity mode on the FCH and RCH may be negotiated at call setup. The diversity mode transmits data on one "spatial mode" using a pair of antennas for each subband.

[00165] The beam-steering mode may be employed on the RACH by user terminals with multiple transmit antennas. A user terminal can estimate the MIMO channel based on the MIMO pilot sent on the BCH. This channel estimate may then be used to perform beam-steering on the RACH for system accesses. The beam-steering mode may also be used for the dedicated transport channels (FCH and RCH). The beam-steering mode may be able to achieve higher received signal-to-noise-and-interference ratio (SNR) at the receiver than the diversity mode by exploiting the gain of the antenna array at the transmitter. In addition, the preamble portion of the PDU may be reduced since the steered reference only includes symbols for a single "steered" antenna. The diversity mode may also be used for the RACH.

[00166] The spatial multiplexing mode may be used for the FCH and RCH to achieve greater throughput, when supported by the channel conditions. The spatial multiplexing and beam-steering modes are reference driven and require closed-loop control for proper operation. As such, a user terminal is allocated resources on both the FCH and RCH to support the spatial multiplexing mode. Up to four spatial channels may be supported on the FCH and RCH (limited by the number of antennas at the access point).

2. Coding and Modulation

[00167] A number of different rates are supported for the transport channels. Each rate is associated with a particular code rate and a particular modulation scheme, which collectively results in a particular spectral efficiency (or data rate). Table 25 lists the various rates supported by the system.

Table 25

Rate Word	Spectral Efficiency (bps/Hz)	Code Rate	Modulation Scheme	Info bits/ OFDM symbol	Code bits/ OFDM symbol
0000	0.0	-	off	-	-
0001	0.25	1/4	BPSK	12	48
0010	0.5	1/2	BPSK	24	48
0011	1.0	1/2	QPSK	48	96
0100	1.5	3/4	QPSK	72	96
0101	2.0	1/2	16 QAM	96	192
0110	2.5	5/8	16 QAM	120	192
0111	3.0	3/4	16 QAM	144	192
1000	3.5	7/12	64 QAM	168	288
1001	4.0	2/3	64 QAM	192	288
1010	4.5	3/4	64 QAM	216	288
1011	5.0	5/6	64 QAM	240	288
1100	5.5	11/16	256 QAM	264	384
1101	6.0	3/4	256 QAM	288	384
1110	6.5	13/16	256 QAM	312	384
1111	7.0	7/8	256 QAM	336	384

[00168] Each common transport channel supports one or more rates and one transmission mode (or possibly more, as may be the case for the RACH). The BCH is transmitted at a fixed rate using the diversity mode. The FCCH may be transmitted at one of four possible rates, as indicated by the FCCH Phy Mode field in the BCH message, using the diversity mode. In one embodiment, the RACH may be transmitted at one of four possible rates, as indicated by the RACH DRI embedded in the preamble of the RACH PDU, and each RACH message is one of four possible sizes. In another embodiment, the RACH is transmitted at a single rate. Table 26 lists the coding, modulation, and transmission parameters and the message sizes supported by each common transport channel.

Table 26 - Parameters for Common Transport Channels

					Message Size

Transport Channel	Spectral Efficiency (bps/Hz)	Code Rate	Modulation Scheme	Transmission Mode	Message Size	
BCH	0.25	1/4	BPSK	Diversity	120 bits	10 OFDM symbols
FCCH	0.25	1/4	BPSK	Diversity	variable	variable
"	0.5	1/2	BPSK	Diversity	variable	variable
"	1.0	1/2	QPSK	Diversity	variable	variable
"	2.0	1/2	16 QAM	Diversity	variable	variable
RACH	0.25	1/4	BPSK	Beam-Steering	96	8
"	0.5	1/2	BPSK	Beam-Steering	96, 192	4, 8
"	1.0	1/2	QPSK	Beam-Steering	96, 192, 384	2, 4, 8
"	2.0	1/2	16 QAM	Beam-Steering	96, 192, 384, 768	1, 2, 4, 8

The FCCH message is variable in size and given in even number of OFDM symbols.

[00169] The FCH and RCH support all of the rates listed in Table 25. Table 27 lists the coding, modulation, and transmission parameters and the message sizes supported by the FCH and RCH.

Table 27 - Parameters for FCH and RCH

					PHY Frame Size		
					code bits	mod symbols	OFDM symbols
0.25 [A]	1/4	BPSK	72	72	144	288	6
0.5	1/2	BPSK	144	144	288	288	6
1.0	1/2	QPSK	288	288	576	288	6
1.5	3/4	QPSK	432	144	576	288	6
2.0	1/2	16 QAM	576	576	1152	288	6
2.5	5/8	16 QAM	720	432	1152	288	6
3.0	3/4	16 QAM	864	288	1152	288	6
3.5	7/12	64 QAM	1008	720	1728	288	6

4.0	2/3	64 QAM	1152	576	1728	288	6
4.5	3/4	64 QAM	1296	432	1728	288	6
5.0	5/6	64 QAM	1440	288	1728	288	6
5.5	11/16	256 QAM	1584	720	2304	288	6
6.0	3/4	256 QAM	1728	576	2304	288	6
6.5	13/16	256 QAM	1872	432	2304	288	6
7.0	7/8	256 QAM	2016	288	2304	288	6

Note A: each rate 1/2 code bit is repeated on two subbands to obtain an effective code rate of 1/4. The parity bits represent redundancy bits introduced by the coding and are used for error correction by the receiver.

[00170] The PHY frame size in Table 27 indicates the number of code bits, modulation symbols, and OFDM symbols for each PHY frame. If 48 data subbands are used for data transmission, then each OFDM symbol includes 48 modulation symbols. For the diversity and beam-steering modes, one symbol stream is transmitted and the PHY frame size corresponds to the single rate employed for this symbol stream. For the spatial multiplexing mode, multiple symbol streams may be sent on multiple spatial channels, and the overall PHY frame size is determined by the sum of the PHY frame sizes for the individual spatial channels. The PHY frame size for each spatial channel is determined by the rate employed for that spatial channel.

[00171] As an example, suppose the MIMO channel is capable of supporting four spatial channels operating at spectral efficiencies of 0.5, 1.5, 4.5, and 5.5 bps/Hz. The four rates selected for the four spatial channels would then be as shown in Table 28.

Table 28 - Example Spatial Multiplexing Transmission

Spatial channel Index	**Spectral Efficiency** (bps/Hz)	**Code Rate**	**Modulation Scheme**	**Payload** (bits/ PHY frame)	**PHY Frame Size**		
					code bits	mod symbols	OFDM symbols
1	0.5	1/2	BPSK	144	288	288	6
2	1.5	3/4	QPSK	432	576	288	6
3	4.5	3/4	64 QAM	1296	1728	288	6
4	5.5	11/16	256 QAM	1584	2304	288	6

The overall PHY frame size is then $144+432+1296+1584=3456$ information bits or $288+576+1728+2304=4896$ code bits. Even though each of the four spatial channels supports a different number of payload bits, the overall PHY frame can be transmitted in 6 OFDM symbols (e.g., 24 μsec, assuming 4 μsec/OFDM symbol).

V. Physical Layer Processing

[00172] **FIG. 7** shows a block diagram of an embodiment of an access point 110x and two user terminals 120x and 120y within the MIMO WLAN system.

[00173] On the downlink, at access point 110x, a transmit (TX) data processor 710 receives traffic data (i.e., information bits) from a data source 708 and signaling and other information from a controller 730 and possibly a scheduler 734. These various types of data may be sent on different transport channels. TX data processor 710 "frames" the data (if necessary), scrambles the framed/unframed data, encodes the scrambled data, interleaves (i.e., reorders) the coded data, and maps the interleaved data into modulation symbols. For simplicity, a "data symbol" refers to a modulation symbol for traffic data, and a "pilot symbol" refers to a modulation symbol for pilot. The scrambling randomizes the data bits. The encoding increases the reliability of the data transmission. The interleaving provides time, frequency, and/or spatial diversity for the code bits. The scrambling, coding, and modulation may be performed based on control signals provided by controller 730 and are described in further detail below. TX data processor 710 provides a stream of modulation symbols for each spatial channel used for data transmission.

[00174] A TX spatial processor 720 receives one or more modulation symbol streams from TX data processor 710 and performs spatial processing on the modulation symbols to provide four streams of transmit symbols, one stream for each transmit antenna. The spatial processing is described in further detail below.

[00175] Each modulator (MOD) 722 receives and processes a respective transmit symbol stream to provide a corresponding stream of OFDM symbols. Each OFDM symbol stream is further processed to provide a corresponding downlink modulated signal. The four downlink modulated signals from modulator 722a through 722d are then transmitted from four antennas 724a through 724d, respectively.

[00176] At each user terminal 120, one or multiple antennas 752 receive the transmitted downlink modulated signals, and each receive antenna provides a received signal to a

respective demodulator (DEMOD) 754. Each demodulator 754 performs processing complementary to that performed at modulator 722 and provides received symbols. A receive (RX) spatial processor 760 then performs spatial processing on the received symbols from all demodulators 754 to provide recovered symbols, which are estimates of the modulation symbols sent by the access point.

[00177] An RX data processor 770 receives and demultiplexes the recovered symbols into their respective transport channels. The recovered symbols for each transport channel may be symbol demapped, deinterleaved, decoded, and descrambled to provide decoded data for that transport channel. The decoded data for each transport channel may include recovered packet data, messages, signaling, and so on, which are provided to a data sink 772 for storage and/or a controller 780 for further processing.

[00178] The processing by access point 110 and terminal 120 for the downlink is described in further detail below. The processing for the uplink may be the same or different from the processing for the downlink.

[00179] For the downlink, at each active user terminal 120, RX spatial processor 760 further estimates the downlink to obtain channel state information (CSI). The CSI may include channel response estimates, received SNRs, and so on. RX data processor 770 may also provide the status of each packet/frame received on the downlink. A controller 780 receives the channel state information and the packet/frame status and determines the feedback information to be sent back to the access point. The feedback information is processed by a TX data processor 790 and a TX spatial processor 792 (if present), conditioned by one or more modulators 754, and transmitted via one or more antennas 752 back to the access point.

[00180] At access point 110, the transmitted uplink signal(s) are received by antennas 724, demodulated by demodulators 722, and processed by an RX spatial processor 740 and an RX data processor 742 in a complementary manner to that performed at the user terminal. The recovered feedback information is then provided to controller 730 and a scheduler 734.

[00181] Scheduler 734 uses the feedback information to perform a number of functions such as (1) selecting a set of user terminals for data transmission on the downlink and uplink, (2) selecting the transmission rate(s) and the transmission mode for each selected user terminal, and (3) assigning the available FCH/RCH resources to the selected terminals. Scheduler 734 and/or controller 730 further uses information (e.g.,

steering vectors) obtained from the uplink transmission for the processing of the downlink transmission, as described in further detail below.

[00182] A number of transmission modes are supported for data transmission on the downlink and uplink. The processing for each of these transmission modes is described in further detail below.

1. Diversity Mode - Transmit Processing

[00183] **FIG. 8A** shows a block diagram of an embodiment of a transmitter unit 800 capable of performing the transmit processing for the diversity mode. Transmitter unit 800 may be used for transmitter portion of the access point and the user terminal.

[00184] Within a TX data processor 710a, a framing unit 808 frames the data for each packet to be transmitted on the FCH or RCH. The framing need not be performed for the other transport channels. The framing may be performed as illustrated in FIG. 6 to generate one or more PHY frames for each user packet. A scrambler 810 then scrambles the framed/unframed data for each transport channel to randomize the data.

[00185] An encoder 812 receives and codes the scrambled data in accordance with a selected coding scheme to provide code bits. A repeat/puncture unit 814 then repeats or punctures (i.e., deletes) some of the code bits to obtain the desired code rate. In an embodiment, encoder 812 is a rate 1/2, constraint length 7, binary convolutional encoder. A code rate of 1/4 may be obtained by repeating each code bit once. Code rates greater than 1/2 may be obtained by deleting some of the code bits from encoder 812. A specific design for framing unit 808, scrambler 810, encoder 812, and repeat/puncture unit 814 is described below.

[00186] An interleaver 818 then interleaves (i.e., reorders) the code bits from unit 814 based on a selected interleaving scheme. In an embodiment, each group of 48 consecutive code bits to be transmitted on a given spatial channel is spread over the 48 data-carrying subbands (or simply, data subbands) to provide frequency diversity. The interleaving is described in further detail below.

[00187] A symbol mapping unit 820 then maps the interleaved data in accordance with a particular modulation scheme to provide modulation symbols. As shown in Table 26, BPSK, 4 QAM, or 16 QAM may be used for the diversity mode, depending on the selected rate. In the diversity mode, the same modulation scheme is used for all data subbands. The symbol mapping may be achieved by (1) grouping sets of B bits to form B-bit values, where $B \geq 1$, and (2) mapping each B-bit value to a point in a signal

constellation corresponding to the selected modulation scheme. Each mapped signal point is a complex value and corresponds to a modulation symbol. Symbol mapping unit 820 provides a stream of modulation symbols to a TX diversity processor 720a.

[00188] In an embodiment, the diversity mode utilizes space-time transmit diversity (STTD) for dual transmit diversity on a per subband basis. STTD supports simultaneous transmission of independent symbol streams on two transmit antennas while maintaining orthogonality at the receiver.

[00189] The STTD scheme operates as follows. Suppose that two modulation symbols, denoted as s_1 and s_2, are to be transmitted on a given subband. The transmitter generates two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2]^T$ and $\underline{\mathbf{x}}_2 = [s_2^* \;\; -s_1^*]^T$, where " * " denotes the complex conjugate and " ^{T} " denotes the transpose. Each vector includes two elements that are to be transmitted from two transmit antennas in one symbol period (i.e., vector $\underline{\mathbf{x}}_1$ is transmitted from two antennas in the first symbol period, and vector $\underline{\mathbf{x}}_2$ is transmitted from two antennas in the next symbol period).

[00190] If the receiver is equipped with a single receive antenna, then the received symbols may be expressed as:

$$r_1 = h_1 s_1 + h_2 s_2 + n_1 \quad \text{, and ,} \qquad \text{Eq (1)}$$

$$r_2 = h_1 s_2^* - h_2 s_1^* + n_2 \quad ,$$

where r_1 and r_2 are two symbols received in two consecutive symbol periods at the receiver;

h_1 and h_2 are the path gains from the two transmit antennas to the receive antenna for the subband under consideration, where the path gains are assumed to be constant over the subband and static over the 2-symbol period; and

n_1 and n_2 are the noise associated with the two received symbols r_1 and r_2, respectively.

[00191] The receiver may then derive estimates of the two transmitted symbols, s_1 and s_2, as follows:

$$\hat{s}_1 = \frac{h_1^* r_1 - h_2 r_2^*}{|h_1|^2 + |h_2|^2} = s_1 + \frac{h_1^* n_1 - h_2 n_2^*}{|h_1|^2 + |h_2|^2} \quad \text{, and} \qquad \text{Eq (2)}$$

$$\hat{s}_2 = \frac{h_2^* r_1 + h_1 r_2^*}{|h_1|^2 + |h_2|^2} = s_2 + \frac{h_2^* n_1 + h_1 n_2^*}{|h_1|^2 + |h_2|^2} \ .$$

[00192] Alternatively, the transmitter may generate two vectors $\underline{\mathbf{x}}_1 = [s_1 \ \ \text{-} s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \ \ s_1^*]^T$ and transmit the two vectors sequentially in two symbol periods from two transmit antennas. The received symbols may then be expressed as:

$$r_1 = h_1 s_1 - h_2 s_2^* + n_1 \quad , \text{ and}$$

$$r_2 = h_1 s_2 + h_2 s_1^* + n_2 \quad .$$

The receiver may then derive estimates of the two transmitted symbols as follows:

$$\hat{s}_1 = \frac{h_1^* r_1 + h_2 r_2^*}{|h_1|^2 + |h_2|^2} = s_1 + \frac{h_1^* n_1 + h_2 n_2}{|h_1|^2 + |h_2|^2} \quad , \text{ and}$$

$$\hat{s}_2 = \frac{-h_2 r_1^* + h_1^* r_2}{|h_1|^2 + |h_2|^2} = s_2 + \frac{h_1^* n_2 - h_2 n_1}{|h_1|^2 + |h_2|^2} \quad .$$

[00193] The above description may be extended for a MIMO-OFDM system with two or more transmit antennas, N_R receive antennas, and multiple subbands. Two transmit antennas are used for any given subband. Suppose that two modulation symbols, denoted as $s_1(k)$ and $s_2(k)$, are to be transmitted on a given subband k. The transmitter generates two vectors $\underline{\mathbf{x}}_1(k) = [s_1(k) \ \ s_2(k)]^T$ and $\underline{\mathbf{x}}_2(k) = [s_2^*(k) \ \ \text{-} s_1^*(k)]^T$ or equivalently two symbol sets $\{x_i(k)\} = \{s_1(k) \ \ s_2^*(k)\}$ and $\{x_j(k)\} = \{s_2(k) \ \ -s_1^*(k)\}$. Each symbol set includes two elements that are to be transmitted sequentially in two symbol periods from a respective transmit antenna on subband k (i.e., symbol set $\{x_i(k)\}$ is transmitted on subband k from antenna i in two symbol periods, and symbol set $\{x_j(k)\}$ is transmitted on subband k from antenna j in the same 2-symbol period).

[00194] The vectors of received symbols at the receive antennas in the two symbol periods may be expressed as:

$$\underline{\mathbf{r}}_1(k) = \underline{\mathbf{h}}_i(k) s_1(k) + \underline{\mathbf{h}}_j(k) s_2(k) + \underline{\mathbf{n}}_1(k) \quad , \text{ and}$$

$$\underline{\mathbf{r}}_2(k) = \underline{\mathbf{h}}_i(k) s_2^*(k) - \underline{\mathbf{h}}_j(k) s_1^*(k) + \underline{\mathbf{n}}_2(k) \quad ,$$

where $\underline{\mathbf{r}}_1(k)$ and $\underline{\mathbf{r}}_2(k)$ are two symbol vectors received in two consecutive symbol periods on subband k at the receiver, with each vector including N_R received symbols for N_R receive antennas;

$\underline{\mathbf{h}}_i(k)$ and $\underline{\mathbf{h}}_j(k)$ are the vectors of path gains from the two transmit antennas i and j to the N_R receive antennas for subband k, with each vector including the channel gains from the associated transmit antenna to each of the N_R receive antennas, where the path gains are assumed to be constant over the subband and static over the 2-symbol period; and

$\underline{\mathbf{n}}_1(k)$ and $\underline{\mathbf{n}}_2(k)$ are noise vectors associated with the two received vectors $\underline{\mathbf{r}}_1(k)$ and $\underline{\mathbf{r}}_2(k)$, respectively.

[00195] The receiver may then derive estimates of the two transmitted symbols, $s_1(k)$ and $s_2(k)$, as follows:

$$\hat{s}_1(k) = \frac{\hat{\underline{\mathbf{h}}}_i^H(k)\underline{\mathbf{r}}_1(k) - \underline{\mathbf{r}}_2^H(k)\hat{\underline{\mathbf{h}}}_j(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2} = s_1(k) + \frac{\hat{\underline{\mathbf{h}}}_i^H(k)\underline{\mathbf{n}}_1(k) - \underline{\mathbf{n}}_2^H(k)\hat{\underline{\mathbf{h}}}_j(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2}, \text{ and}$$

$$\hat{s}_2(k) = \frac{\hat{\underline{\mathbf{h}}}_j^H(k)\underline{\mathbf{r}}_1(k) + \underline{\mathbf{r}}_2^H(k)\hat{\underline{\mathbf{h}}}_i(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2} = s_2(k) + \frac{\hat{\underline{\mathbf{h}}}_j^H(k)\underline{\mathbf{n}}_1(k) + \underline{\mathbf{n}}_2^H(k)\hat{\underline{\mathbf{h}}}_i(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2}.$$

[00196] Alternatively, the transmitter may generate two symbol sets $\{x_i(k)\} = \{s_1(k) \;\; s_2(k)\}$ and $\{x_j(k)\} = \{-s_2^*(k) \;\; s_1^*(k)\}$ and transmit these two symbol sets from two transmit antennas i and j. The vectors of received symbols may then be expressed as:

$$\underline{\mathbf{r}}_1(k) = \underline{\mathbf{h}}_i(k)s_1(k) - \underline{\mathbf{h}}_j(k)s_2^*(k) + \underline{\mathbf{n}}_1(k) \text{ , and}$$

$$\underline{\mathbf{r}}_2(k) = \underline{\mathbf{h}}_i(k)s_2(k) + \underline{\mathbf{h}}_j(k)s_1^*(k) + \underline{\mathbf{n}}_2(k) \text{ .}$$

The receiver may then derive estimates of the two transmitted symbols as follows:

$$\hat{s}_1(k) = \frac{\hat{\underline{\mathbf{h}}}_i^H(k)\underline{\mathbf{r}}_1(k) + \underline{\mathbf{r}}_2^H(k)\hat{\underline{\mathbf{h}}}_j(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2} = s_1(k) + \frac{\hat{\underline{\mathbf{h}}}_i^H(k)\underline{\mathbf{n}}_1(k) + \underline{\mathbf{n}}_2^H(k)\hat{\underline{\mathbf{h}}}_j(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2}, \text{ and}$$

$$\hat{s}_2(k) = \frac{\hat{\underline{\mathbf{h}}}_i^H(k)\underline{\mathbf{r}}_2(k) - \underline{\mathbf{r}}_1^H(k)\hat{\underline{\mathbf{h}}}_j(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2} = s_2(k) + \frac{\hat{\underline{\mathbf{h}}}_i^H(k)\underline{\mathbf{n}}_2(k) - \underline{\mathbf{n}}_1^H(k)\hat{\underline{\mathbf{h}}}_j(k)}{\|\hat{\underline{\mathbf{h}}}_i(k)\|^2 + \|\hat{\underline{\mathbf{h}}}_j(k)\|^2}.$$

[00197] The STTD scheme is described by S.M. Alamouti in a paper entitled "A Simple Transmit Diversity Technique for Wireless Communications," IEEE Journal on Selected Areas in Communications, Vol. 16, No. 8, October 1998, pgs. 1451-1458. The STTD scheme is also described in commonly assigned U.S. Patent Application Serial No. 09/737,602, entitled "Method and System for Increased Bandwidth Efficiency in Multiple Input - Multiple Output Channels," filed January 5, 2001, and U.S. Patent

Application Serial No. 10/179,439, entitled "Diversity Transmission Modes for MIMO OFDM Communication Systems," filed June 24, 2002.

[00198] The STTD scheme effectively transmits one modulation symbol per subband over two transmit antennas in each symbol period. However, the STTD scheme distributes the information in each modulation symbol over two successive OFDM symbols. Thus, the symbol recovery at the receiver is performed based on two consecutive received OFDM symbols.

[00199] The STTD scheme utilizes one pair of transmit antennas for each data subband. Since the access point includes four transmit antennas, each antenna may be selected for use for half of the 48 data subbands. Table 29 lists an exemplary subband-antenna assignment scheme for the STTD scheme.

Table 29

Subband Indices	TX Ant	Bit Index	Subband Indices	TX Ant	Bit Index	Subband Indices	TX Ant	Bit Index	Subband Indices	TX Ant	Bit Index
-	-	-	-13	1,2	26	1	3,4	1	15	1,2	33
-26	1,2	0	-12	3,4	32	2	1,2	7	16	2,4	39
-25	3,4	6	-11	1,3	38	3	2,4	13	17	1,3	45
-24	1,3	12	-10	2,4	44	4	1,3	19	18	2,3	5
-23	2,4	18	-9	1,4	4	5	2,3	25	19	1,4	11
-22	1,4	24	-8	2,3	10	6	1,4	31	20	3,4	17
-21	1	**P0**	-7	2	**P1**	7	3	**P2**	21	4	**P3**
-20	2,3	30	-6	1,2	16	8	3,4	37	22	1,2	23
-19	1,2	36	-5	3,4	22	9	1,2	43	23	2,4	29
-18	3,4	42	-4	1,3	28	10	2,4	3	24	1,3	35
-17	1,3	2	-3	2,4	34	11	1,3	9	25	2,3	41
-16	2,4	8	-2	1,4	40	12	2,3	15	26	1,4	47
-15	1,4	14	-1	2,3	46	13	1,4	21	-	-	-
-14	2,3	20	0	-	-	14	3,4	27	-	-	-

[00200] As shown in Table 29, transmit antennas 1 and 2 are used for subbands with indices -26, -19, -13, and so on, transmit antennas 2 and 4 are used for subbands with indices -25, -18, -12, and so on, transmit antennas 1 and 3 are used for subbands with

indices -24, -17, -11, and so on. There are six different antenna pairings with four transmit antennas. Each of the six antenna pairings is used for 8 data subbands, which are spaced approximately uniformly across the 48 data subbands. The antenna pairing to subband assignment is such that different antennas are used for adjacent subbands, which may provide greater frequency and spatial diversity. For example, antennas 1 and 2 are used for subband -26, and antennas 3 and 4 are used for subband -25.

[00201] The antenna-subband assignment in Table 29 is also such that all four transmit antennas are used for each code bit for the lowest rate of 1/4, which may maximize spatial diversity. For rate 1/4, each code bit is repeated and sent on two subbands (which is also referred to as dual subband repeat coding). The two subbands used for each code bit are mapped to different antenna pairs so that all four antennas are used to transmit that code bit. For example, bit indices 0 and 1 in Table 29 correspond to the same code bit for the diversity mode, where the bit with index 0 is transmitted from antennas 1 and 2 on subband -26 and the bit with index 1 is transmitted from antennas 3 and 4 on subband 1. As another example, bit indices 2 and 3 in Table 29 correspond to the same code bit, where the bit with index 2 is transmitted from antennas 1 and 3 in subband -17 and the bit with index 3 is transmitted from antennas 2 and 4 in subband 10.

[00202] The system may support other transmit diversity schemes, and this is within the scope of the invention. For example, the system may support a space-frequency transmit diversity (SFTD) that can achieve space and frequency diversity on a per-subband-pair basis. An exemplary SFTD scheme operates as follows. Suppose that two modulation symbols, denoted as $s(k)$ and $s(k+1)$, are generated and mapped to two adjacent subbands of an OFDM symbol. For SFTD, the transmitter would transmit symbols $s(k)$ and $s(k+1)$ from two antennas on subband k and would transmit symbols $s^*(k+1)$ and $-s^*(k)$ from the same two antennas on subband $k+1$. Adjacent subbands are used for the pair of modulation symbols because the channel response is assumed to be constant for the transmission of the two pairs of symbols. The processing at the receiver to recover the modulation symbols is the same as for the STTD scheme, except that the received symbols for two subbands instead of two OFDM symbol periods are processed.

[00203] **FIG. 8B** shows a block diagram of an embodiment of a TX diversity processor 720a capable of implementing the STTD scheme for the diversity mode.

[00204] Within TX diversity processor 720a, a demultiplexer 832 receives and demultiplexes the stream of modulation symbols $s(n)$ from TX data processor 710a into 48 substreams, denoted as $s_1(n)$ through $s_k(n)$, for the 48 data subbands. Each modulation symbol substream includes one modulation symbol for each symbol period, which corresponds to a symbol rate of $(T_{OFDM})^{-1}$, where T_{OFDM} is the duration of one OFDM symbol. Each modulation symbol substream is provided to a respective TX subband diversity processor 840.

[00205] Within each TX subband diversity processor 840, a demultiplexer 842 demultiplexes the modulation symbols for the subband into two symbol sequences, with each sequence having a symbol rate of $(2T_{OFDM})^{-1}$. A space-time encoder 850 receives the two modulation symbol sequences and, for each 2-symbol period, uses two symbols s_1 and s_2 in the two sequences to form two symbol sets $\{x_i\} = \{s_1 \;\; s_2^*\}$ and $\{x_j\} = \{s_2 \;\; -s_1^*\}$ for two transmit antennas. Each symbol set includes two symbols, one symbol from each of the two sequences. Symbol set $\{x_i\}$ is generated by providing symbol s_1 first and symbol s_2^* next, where s_1 is obtained via a switch 856a and s_2^* is obtained by taking the conjugate of s_2 with a unit 852a and delaying the conjugated symbol by one symbol period with a delay unit 854a. Similarly, symbol set $\{x_j\}$ is generated by providing symbol s_2 first and symbol $-s_1^*$ next, where s_2 is obtained via a switch 856b and $-s_1^*$ is obtained by taking the negative conjugate of s_1 with a unit 852b and delaying the negative conjugated symbol by one symbol period with a delay unit 854b. The two symbol sets $\{x_i\}$ and $\{x_j\}$ are to be transmitted from two antennas i and j assigned to the subband, as indicated in Table 29. Space-time encoder 850 provides the first symbol set $\{x_i\} = \{s_1 \;\; s_2^*\}$ to a buffer/multiplexer 870 for the first transmit antenna i and the second symbol set $\{x_j\} = \{s_2 \;\; -s_1^*\}$ to another buffer/multiplexer 870 for the second transmit antenna j. The two symbols provided by space-time encoder 850 for each symbol period are referred to as STTD symbols.

[00206] Buffers/multiplexers 870a through 870d are used to buffer and multiplex the STTD symbols from all diversity processors 840. Each buffer/multiplexer 870 receives pilot symbols and STTD symbols from the appropriate TX subband diversity processors 840, as determined by Table 29. For example, buffer/multiplexer 870a receives modulation symbols for subbands -26, -24, -22, -19, and so on (i.e., all subbands mapped to antenna 1), buffer/multiplexer 870b receives modulation symbols for subbands -26, -23, -20, -19, and so on (i.e., all subbands mapped to antenna 2), buffer/multiplexer 870c receives modulation symbols for subbands -25, -24, -20, -18, and so on (i.e., all subbands mapped to antenna 3), and buffer/multiplexer 870d receives modulation symbols for subbands -25, -23, -22, -18, and so on (i.e., all subbands mapped to antenna 4).

[00207] Each buffer/multiplexer 870 then, for each symbol period, multiplexes four pilots, 24 STTD symbols, and 36 zeros for the four pilot subbands, 24 data subbands, and 36 unused subbands, respectively, to form a sequence of 64 transmit symbols for the 64 total subbands. Although there are a total of 48 data subbands, only 24 subbands are used for each transmit antenna for the diversity mode, and the effective total number of unused subbands for each antenna is thus 36 instead of 12. Each transmit symbol is a complex value (which may be zero for an unused subband) that is sent on one subband in one symbol period. Each buffer/multiplexer 870 provides a stream of transmit symbols $x_i(n)$ for one transmit antenna. Each transmit symbol stream comprises concatenated sequences of 64 transmit symbols, one sequence for each symbol period. Referring back to FIG. 8A, TX diversity processor 720a provides four transmit symbol streams, $x_1(n)$ through $x_4(n)$, to four OFDM modulators 722a through 722d.

[00208] **FIG. 8C** shows a block diagram of an embodiment of an OFDM modulator 722x, which may be used for each of OFDM modulators 722a through 722d in FIG. 8A. Within OFDM modulator 722x, an inverse fast Fourier transform (IFFT) unit 852 receives a stream of transmit symbol, $x_i(n)$, and converts each sequence of 64 transmit symbols into its time-domain representation (which is referred to as a transformed symbol) using a 64-point inverse fast Fourier transform. Each transformed symbol comprises 64 time-domain samples corresponding to the 64 total subbands.

[00209] For each transformed symbol, cyclic prefix generator 854 repeats a portion of the transformed symbol to form a corresponding OFDM symbol. As noted above, one

of two different cyclic prefix lengths may be used. The cyclic prefix for the BCH is fixed and is 800 nsec. The cyclic prefix for all other transport channels is selectable (either 400 nsec or 800 nsec) and indicated by the Cyclic Prefix Duration field of the BCH message. For a system with a bandwidth of 20 MHz, a sample period of 50 nsec, and 64 subbands, each transformed symbol has a duration of 3.2 msec (or 64×50 nsec) and each OFDM symbol has a duration of either 3.6 msec or 4.0 msec depending on whether the 400 nsec or 800 nsec cyclic prefix is used for the OFDM symbol.

[00210] **FIG. 8D** illustrates an OFDM symbol. The OFDM symbol is composed of two parts: a cyclic prefix having a duration of 400 or 800 nsec (8 or 16 samples) and a transformed symbol with a duration of 3.2 μsec (64 samples). The cyclic prefix is a copy of the last 8 or 16 samples (i.e., a cyclic continuation) of the transformed symbol and is inserted in front of the transformed symbol. The cyclic prefix ensures that the OFDM symbol retains its orthogonal property in the presence of multipath delay spread, thereby improving performance against deleterious path effects such as multipath and channel dispersion caused by frequency selective fading.

[00211] Cyclic prefix generator 854 provides a stream of OFDM symbols to a transmitter (TMTR) 856. Transmitter 856 converts the OFDM symbol stream into one or more analog signals, and further amplifies, filters, and frequency upconverts the analog signals to generate a modulated signal suitable for transmission from an associated antenna.

[00212] The baseband waveform for an OFDM symbol may be expressed as:

$$x_n(t) = \sum_{k=-N_{ST}/2,k\neq 0}^{N_{ST}/2} c_n(k)\Psi_n(k,t) \ , \qquad \text{Eq (3)}$$

where n denotes the symbol period (i.e., the OFDM symbol index);

k denotes the subband index;

N_{ST} is the number of pilot and data subbands;

$c_n(k)$ denotes the symbol transmitted on subband k of symbol period n; and

$$\Psi_n(k,t) = \begin{cases} e^{j2\pi k\Delta f(t-T_{CP}-nT_S)} & \text{, for } nT_S \le t \le (n+1)T_S \\ 0 & \text{, otherwise} \end{cases} \ , \qquad \text{Eq (4)}$$

where T_{CP} is the cyclic prefix duration;

T_S is the OFDM symbol duration; and

Δf is the bandwidth of each subband.

2. Spatial Multiplexing Mode - Transmit Processing

[00213] **FIG. 9A** shows a block diagram of a transmitter unit 900 capable of performing the transmit processing for the spatial multiplexing mode. Transmitter unit 900 is another embodiment of the transmitter portion of the access point and the user terminal. For the spatial multiplexing mode, again assuming that four transmit antennas and four receive antennas are available, data may be transmitted on up to four spatial channels. A different rate may be used for each spatial channel depending on its transmission capacity. Each rate is associated with a particular code rate and modulation scheme, as shown in Table 25. In the following description it is assumed that N_E spatial channels are selected for use for data transmission, where $N_E \leq N_S \leq \min\{N_T, N_R\}$.

[00214] Within a TX data processor 710b, framing unit 808 frames the data for each FCH/RCH packet to generate one or more PHY frames for the packet. Each PHY frame includes the number of data bits that may be transmitted in all N_E spatial channels within 6 OFDM symbols. Scrambler 810 scrambles the data for each transport channel. Encoder 812 receives and codes the scrambled data in accordance with a selected coding scheme to provide code bits. In an embodiment, a common coding scheme is used to code the data for all N_E spatial channels, and different code rates for different spatial channels are obtained by puncturing the code bits with different puncturing patterns. Puncture unit 814 thus punctures the code bits to obtain the desired code rate for each spatial channel. The puncturing for the spatial multiplexing mode is described in further detail below.

[00215] A demultiplexer 816 receives and demultiplexes the code bits from puncture unit 814 to provide N_E code bit streams for the N_E spatial channels selected for use. Each code bit stream is provided to a respective interleaver 818, which interleaves the code bits in the stream across the 48 data subbands. The coding and interleaving for the spatial multiplexing mode are described in further detail below. The interleaved data from each interleaver 818 is provided to a respective symbol mapping unit 820.

[00216] In the spatial multiplexing mode, up to four different rates may be used for the four spatial channels, depending on the received SNRs achieved for these spatial channels. Each rate is associated with a particular modulation scheme, as shown in Table 25. Each symbol mapping unit 820 maps the interleaved data in accordance with a particular modulation scheme selected for the associated spatial channel to provide modulation symbols. If all four spatial channels are selected for use, then symbol

mapping units 820a through 820d provide four streams of modulation symbols for the four spatial channels to a TX spatial processor 720b.

[00217] TX spatial processor 720b performs spatial processing for the spatial multiplexing mode. For simplicity, the following description assumes that four transmit antennas, four receive antennas, and 48 data subbands are used for data transmission. The data subband indices are given by the set *K*, where $K = \pm\{1, ..., 6, 8 ..., 20, 22, ... 26\}$ for the OFDM subband structure described above.

[00218] The model for a MIMO-OFDM system may be expressed as:

$$\underline{\mathbf{r}}(k) = \underline{\mathbf{H}}(k)\underline{\mathbf{x}}(k) + \underline{\mathbf{n}}(k) \quad \text{, for } k \in K \text{ ,} \qquad \text{Eq (5)}$$

where $\underline{\mathbf{r}}(k)$ is a "receive" vector with four entries for the symbols received via the four receive antennas for subband *k* (i.e., $\underline{\mathbf{r}}(k) = [r_1(k) \ r_2(k) \ r_3(k) \ r_4(k)]^T$);

$\underline{\mathbf{x}}(k)$ is a "transmit" vector with four entries for the symbols transmitted from the four transmit antennas for subband *k* (i.e., $\underline{\mathbf{x}}(k) = [x_1(k) \ x_2(k) \ x_3(k) \ x_4(k)]^T$);

$\underline{\mathbf{H}}(k)$ is an $(N_R \times N_T)$ channel response matrix for subband *k*; and

$\underline{\mathbf{n}}(k)$ is a vector of additive white Gaussian noise (AWGN) for subband *k*.

The noise vector $\underline{\mathbf{n}}(k)$ is assumed to have components with zero mean and a covariance matrix of $\underline{\Lambda}_n = \sigma^2\underline{\mathbf{I}}$, where $\underline{\mathbf{I}}$ is the identity matrix and σ^2 is the noise variance.

[00219] The channel response matrix $\underline{\mathbf{H}}(k)$ for subband *k* may be expressed as:

$$\underline{\mathbf{H}}(k) = \begin{bmatrix} h_{1,1}(k) & h_{1,2}(k) & h_{1,3}(k) & h_{1,4}(k) \\ h_{2,1}(k) & h_{2,2}(k) & h_{2,3}(k) & h_{2,4}(k) \\ h_{3,1}(k) & h_{3,2}(k) & h_{3,3}(k) & h_{3,4}(k) \\ h_{4,1}(k) & h_{4,2}(k) & h_{4,3}(k) & h_{4,4}(k) \end{bmatrix} \text{, for } k \in K \text{ ,} \qquad \text{Eq (6)}$$

where entry $h_{ij}(k)$, for $i \in \{1, 2, 3, 4\}$ and $j \in \{1, 2, 3, 4\}$, is the coupling (i.e., complex gain) between transmit antenna *i* and receive antenna *j* for subband *k*. For simplicity, it is assumed that the channel response matrices $\underline{\mathbf{H}}(k)$, for $k \in K$, are known or can be ascertained by both the transmitter and receiver.

[00220] The channel response matrix $\underline{\mathbf{H}}(k)$ for each subband may be "diagonalized" to obtain the N_S eigenmodes for that subband. This can be achieved by performing eigenvalue decomposition on the correlation matrix of $\underline{\mathbf{H}}(k)$, which is

$\underline{\mathbf{R}}(k) = \underline{\mathbf{H}}^H(k)\underline{\mathbf{H}}(k)$, where $\underline{\mathbf{H}}^H(k)$ denotes the conjugate transpose of $\underline{\mathbf{H}}(k)$. The eigenvalue decomposition of the correlation matrix $\underline{\mathbf{R}}(k)$ may be expressed as:

$$\underline{\mathbf{R}}(k) = \underline{\mathbf{V}}(k)\underline{\mathbf{D}}(k)\underline{\mathbf{V}}^H(k) \quad \text{, for } k \in K \text{ ,} \qquad \text{Eq (7)}$$

where $\underline{\mathbf{V}}(k)$ is an $(N_T \times N_T)$ unitary matrix whose columns are eigenvectors of $\underline{\mathbf{R}}(k)$ (i.e., $\underline{\mathbf{V}}(k) = [\underline{\mathbf{v}}_1(k)\ \underline{\mathbf{v}}_2(k)\ \underline{\mathbf{v}}_3(k)\ \underline{\mathbf{v}}_4(k)]$, where each $\underline{\mathbf{v}}_i(k)$ is an eigenvector for one eigenmode); and

$\underline{\mathbf{D}}(k)$ is an $(N_T \times N_T)$ diagonal matrix of eigenvalues of $\underline{\mathbf{R}}(k)$.

A unitary matrix is characterized by the property $\underline{\mathbf{M}}^H \underline{\mathbf{M}} = \underline{\mathbf{I}}$. Eigenvectors $\underline{\mathbf{v}}_i(k)$, for $i \in \{1, 2, 3, 4\}$, are also referred to as transmit steering vectors for each of the spatial channels.

[00221] The channel response matrix $\underline{\mathbf{H}}(k)$ may also be diagonalized using singular value decomposition, which may be expressed as:

$$\underline{\mathbf{H}}(k) = \underline{\mathbf{U}}(k)\underline{\Sigma}(k)\underline{\mathbf{V}}^H(k) \quad \text{, for } k \in K \text{ ,} \qquad \text{Eq (8)}$$

where $\underline{\mathbf{V}}(k)$ is a matrix whose columns are right eigenvectors of $\underline{\mathbf{H}}(k)$;

$\underline{\Sigma}(k)$ is a diagonal matrix containing singular values of $\underline{\mathbf{H}}(k)$, which are positive square roots of the diagonal elements of $\underline{\mathbf{D}}(k)$, the eigenvalues of $\underline{\mathbf{R}}(k)$; and

$\underline{\mathbf{U}}(k)$ is a matrix whose columns are left eigenvectors of $\underline{\mathbf{H}}(k)$.

Singular value decomposition is described by Gilbert Strang in a book entitled "Linear Algebra and Its Applications," Second Edition, Academic Press, 1980. As shown in equations (7) and (8), the columns of the matrix $\underline{\mathbf{V}}(k)$ are eigenvectors of $\underline{\mathbf{R}}(k)$ as well as right eigenvectors of $\underline{\mathbf{H}}(k)$. The columns of the matrix $\underline{\mathbf{U}}(k)$ are eigenvectors of $\underline{\mathbf{H}}(k)\underline{\mathbf{H}}^H(k)$ as well as left eigenvectors of $\underline{\mathbf{H}}(k)$.

[00222] The diagonal matrix $\underline{\mathbf{D}}(k)$ for each subband contains non-negative real values along the diagonal and zeros everywhere else. The eigenvalues of $\underline{\mathbf{R}}(k)$ are denoted as $\{\lambda_1(k),\ \lambda_2(k),\ \lambda_3(k),\ \lambda_4(k)\}$ or $\{\lambda_i(k)\}$ for $i \in \{1, 2, 3, 4\}$.

[00223] The eigenvalue decomposition may be performed independently for the channel response matrix $\underline{\mathbf{H}}(k)$ for each of the 48 data subbands to determine the four eigenmodes for that subband (assuming that each matrix $\underline{\mathbf{H}}(k)$ is full rank). The four

eigenvalues for each diagonal matrix $\underline{\mathbf{D}}(k)$ may be ordered such that $\{\lambda_1(k) \geq \lambda_2(k) \geq \lambda_3(k) \geq \lambda_4(k)\}$, where $\lambda_1(k)$ is the largest eigenvalue and $\lambda_4(k)$ is the smallest eigenvalue for subband k. When the eigenvalues for each diagonal matrix $\underline{\mathbf{D}}(k)$ are ordered, the eigenvectors (or columns) of the associated matrix $\underline{\mathbf{V}}(k)$ are also ordered correspondingly.

[00224] A "wideband" eigenmode may be defined as the set of same-order eigenmodes of all subbands after the ordering (i.e., wideband eigenmode m includes eigenmodes m of all subbands). Each wideband eigenmode is associated with a respective set of eigenvectors for all of the subbands. The "principal" wideband eigenmode is the one associated with the largest singular value in each of the matrices $\hat{\underline{\Sigma}}(k)$ after the ordering.

[00225] A vector $\underline{\mathbf{d}}^m$ may then be formed to include the m-th rank eigenvalue for all 48 data subbands. This vector $\underline{\mathbf{d}}^m$ may be expressed as:

$$\underline{\mathbf{d}}^m = [\lambda_m(-26) \ \ldots \ \lambda_m(-22) \ \ldots \lambda_m(22) \ \ldots \ \lambda_m(26)] \ , \ \text{for } m = \{1,2,3,4\} . \qquad \text{Eq (9)}$$

The vector $\underline{\mathbf{d}}^1$ includes the eigenvalues for the best or principal wideband eigenmode. For a MIMO-OFDM system with four transmit antennas and four receive antennas (i.e., a 4×4 system), there are up four wideband eigenmodes.

[00226] If the noise variance σ^2 at the receiver is constant across the operating band and known to the transmitter, then the received SNR for each subband of each wideband eigenmode may be determined by dividing the eigenvalue $\lambda_m(k)$ by the noise variance σ^2. For simplicity, the noise variance can be assumed to be equal to one (i.e., $\sigma^2 = 1$).

[00227] For the spatial multiplexing mode, the total transmit power P_{total} available to the transmitter may be distributed to the wideband eigenmodes based on various power allocation schemes. In one scheme, the total transmit power P_{total} is distributed uniformly to all four wideband eigenmodes such that $P_m = P_{total}/4$, where P_m is the transmit power allocated to wideband eigenmode m. In another scheme, the total transmit power P_{total} is distributed to the four wideband eigenmodes using a water-filling procedure.

[00228] The water-filling procedure distributes power such that the wideband eigenmodes with higher power gains receive greater fractions of the total transmit

power. The amount of transmit power allocated to a given wideband eigenmode is determined by its received SNR, which in turn is dependent on the power gains (or eigenvalues) for all of the subbands of that wideband eigenmode. The water-filling procedure may allocate zero transmit power to wideband eigenmodes with sufficiently poor received SNRs. The water-filling procedure receives $\underline{\boldsymbol{\beta}} = \{\beta_1, \beta_2, \beta_3, \beta_4\}$ for the four wideband eigenmodes, where β_m is a normalization factor for wideband eigenmode m and may be expressed as:

$$\beta_m = \frac{1}{\sum_{k \in K} \lambda_m^{-1}(k)} \,, \text{ for } m = \{1, 2, 3, 4\} . \qquad \text{Eq (10)}$$

The normalization factor β_m keeps the transmit power allocated to wideband eigenmode m invariant after channel inversion is applied, as described below. As shown in equation (10), the normalization factor β_m can be derived based on the eigenvalues in the vector $\underline{\mathbf{d}}^m$ and with the assumption of the noise variance being equal to one (i.e., $\sigma^2 = 1$).

[00229] The water-filling procedure then determines the fraction α_m of the total transmit power to allocate to each wideband eigenmode based on the set $\underline{\boldsymbol{\beta}}$ such that spectral efficiency or some other criterion is optimized. The transmit power allocated to wideband eigenmode m by the water-filling procedure may be expressed as:

$$P_m = \alpha_m P_{total} \,, \text{ for } m = \{1, 2, 3, 4\} . \qquad \text{Eq (11)}$$

The power allocations for the four wideband eigenmodes may be given by $\underline{\boldsymbol{\alpha}} = \{\alpha_1, \alpha_2, \alpha_3, \alpha_4\}$, where $\sum_{m=1}^{4} \alpha_m = 1$ and $\sum_{m=1}^{4} P_m = P_{total}$. The spatial multiplexing mode may be selected for use if more than one value in set $\underline{\boldsymbol{\alpha}}$ is non-zero.

[00230] The procedure for performing water-filling is known in the art and not described herein. One reference that describes water-filling is "Information Theory and Reliable Communication," by Robert G. Gallager, John Wiley and Sons, 1968, which is incorporated herein by reference.

[00231] For the spatial multiplexing mode, the rate for each spatial channel or wideband eigenmode may be selected based on the received SNR achieved by that spatial channel/wideband eigenmode with its allocated transmit power of P_m. For simplicity,

the following description assumes data transmission on the wideband eigenmodes. The received SNR for each wideband eigenmode may be expressed as:

$$\gamma_m = \frac{P_m \beta_m}{\sigma^2} \text{ , for } m = \{1, 2, 3, 4\} . \qquad \text{Eq (12)}$$

In one embodiment, the rate for each wideband eigenmode is determined based on a table that includes the rates supported by the system and a range of SNRs for each rate. This table may be obtained by computer simulation, empirical measurements, and so on. The particular rate to use for each wideband eigenmode is the rate in the table with a range of SNRs covering the received SNR for the wideband eigenmode. In another embodiment, the rate for each wideband eigenmode is selected based on (1) the received SNR for the wideband eigenmode, (2) an SNR offset used to account for estimation error, variability in the MIMO channel, and other factors, and (3) a table of supported rates and their required SNRs. For this embodiment, an average received SNR for each wideband eigenmode is first computed as described above or as an average of the received SNRs (in units of dBs) for all of the subbands of the wideband eigenmode. In any case, an operating SNR is next computed as the sum of the received SNR and the SNR offset (where both are given in units of dBs). The operating SNR is then compared against the required SNR for each of the rates supported by the system. The highest rate in the table with a required SNR that is less than or equal to the operating SNR is then selected for the wideband eigenmode. The rate for the transmit diversity mode and the beam-steering mode may also be determined in similar manner.

[00232] The transmit power P_m allocated for each wideband eigenmode may be distributed across the 48 data subbands of that wideband eigenmode such that the received SNRs for all subbands are approximately equal. This non-uniform allocation of power across the subbands is referred to as channel inversion. The transmit power $P_m(k)$ allocated to each subband may be expressed as:

$$P_m(k) = \frac{\beta_m P_m}{\lambda_m(k)} \text{ , for } k \in K \text{ and } m = \{1, 2, 3, 4\} , \qquad \text{Eq (13)}$$

where β_m is given in equation (10).

[00233] As shown in equation (13), the transmit power P_m is distributed non-uniformly across the data subbands based on their channel power gains, which is given by the eigenvalues $\lambda_m(k)$, for $k \in K$. The power distribution is such that approximately equal

received SNRs are achieved at the receiver for all data subbands of each wideband eigenmode. This channel inversion is performed independently for each of the four wideband eigenmodes. The channel inversion per wideband eigenmode is described in further detail in commonly assigned U.S. Patent Application Serial No. 10/229,209, entitled "Coded MIMO Systems with Selective Channel Inversion Applied Per Eigenmode," filed August 27, 2002.

[00234] The channel inversion may be performed in various manners. For full channel inversion, all data subbands are used for data transmission if a wideband eigenmode is selected for use. For selective channel inversion, all or a subset of the available data subbands may be selected for use for each wideband eigenmode. The selective channel inversion discards poor subbands with received SNR below a particular threshold and performs channel inversion on only the selected subbands. Selective channel inversion for each wideband eigenmode is also described in commonly assigned U.S. Patent Application Serial No. 10/229,209, entitled "Coded MIMO Systems with Selective Channel Inversion Applied Per Eigenmode," filed August 27, 2002. For simplicity, the following description assumes that full channel inversion is performed for each wideband eigenmode selected for use.

[00235] The gain to use for each subband of each wideband eigenmode may be determined based on the transmit power $P_m(k)$ allocated to that subband. The gain $g_m(k)$ for each data subband may be expressed as:

$$g_m(k) = \sqrt{P_m(k)} \text{ , for } k \in K \text{ and } m = \{1, 2, 3, 4\} . \qquad \text{Eq (14)}$$

A diagonal gain matrix $\underline{\mathbf{G}}(k)$ may be defined for each subband. This matrix $\underline{\mathbf{G}}(k)$ includes the gains for the four eigenmodes for subband k along the diagonal, and may be expressed as: $\underline{\mathbf{G}}(k) = \text{diag}\,[g_1(k),\ g_2(k),\ g_3(k),\ g_4(k)]$.

[00236] For the spatial multiplexing mode, the transmit vector $\underline{\mathbf{x}}(k)$ for each data subband may be expressed as:

$$\underline{\mathbf{x}}(k) = \underline{\mathbf{V}}(k)\underline{\mathbf{G}}(k)\underline{\mathbf{s}}(k) \text{ , for } k \in K , \qquad \text{Eq (15)}$$

where

$$\underline{\mathbf{s}}(k) = [s_1(k)\ \ s_2(k)\ \ s_3(k)\ \ s_4(k)]^T \text{ , and}$$

$$\underline{\mathbf{x}}(k) = [x_1(k)\ \ x_2(k)\ \ x_3(k)\ \ x_4(k)]^T .$$

The vector $\underline{\mathbf{s}}(k)$ includes four modulation symbols to be transmitted on the four eigenmodes for subband k, and the vector $\underline{\mathbf{x}}(k)$ includes four transmit symbols to be transmitted from the four antennas for subband k. For simplicity, equation (15) does not include the correction factors used to account for differences between the transmit/receive chains at the access point and the user terminal, which are described in detail below.

[00237] **FIG. 9B** shows a block diagram of an embodiment of TX spatial processor 720b capable of performing spatial processing for the spatial multiplexing mode. For simplicity, the following description assumes that all four wideband eigenmodes are selected for use. However, less than four wideband eigenmodes may also be selected for use.

[00238] Within processor 720b, a demultiplexer 932 receives the four modulation symbol streams (denoted as $s_1(n)$ through $s_4(n)$) to be transmitted on the four wideband eigenmodes, demultiplexes each stream into 48 substreams for the 48 data subbands, and provides four modulation symbol substreams for each data subband to a respective TX subband spatial processor 940. Each processor 940 performs the processing shown in equation (15) for one subband.

[00239] Within each TX subband spatial processor 940, the four modulation symbol substreams (denoted as $s_1(k)$ through $s_4(k)$) are provided to four multipliers 942a through 942d, which also receive the gains $g_1(k)$, $g_2(k)$, $g_3(k)$, and $g_4(k)$ for the four eigenmodes of the associated subband. Each gain $g_m(k)$ may be determined based on the transmit power $P_m(k)$ allocated to that subband/eigenmode, as shown in equation (14). Each multiplier 942 scales its modulation symbols with its gain $g_m(k)$ to provide scaled modulation symbols. Multipliers 942a through 942d provide four scaled modulation symbol substreams to four beam-formers 950a through 950d, respectively.

[00240] Each beam-former 950 performs beam-forming to transmit one symbol substream on one eigenmode of one subband. Each beam-former 950 receives one symbol substream $s_m(k)$ and one eigenvector $\underline{\mathbf{v}}_m(k)$ for the associated eigenmode. In particular, beam-former 950a receives eigenvector $\underline{\mathbf{v}}_1(k)$ for the first eigenmode, beam-former 950b receives eigenvector $\underline{\mathbf{v}}_2(k)$ for the second eigenmode, and so on. The beam-forming is performed using the eigenvector for the associated eigenmode.

[00241] Within each beam-former 950, the scaled modulation symbols are provided to four multipliers 952a through 952d, which also receive four elements, $v_{m,1}(k)$, $v_{m,2}(k)$, $v_{m,3}(k)$, and $v_{m,4}(k)$, of eigenvector $\underline{\mathbf{v}}_m(k)$ for the associated eigenmode. Each multiplier 952 then multiplies the scaled modulation symbols with its eigenvector value $v_{m,j}(k)$ to provide "beam-formed" symbols. Multipliers 952a through 952d provide four beam-formed symbol substreams (which are to be transmitted from four antennas) to summers 960a through 960d, respectively.

[00242] Each summer 960 receives and sums four beam-formed symbols for the four eigenmodes for each symbol period to provide a preconditioned symbol for an associated transmit antenna. Summers 960a through 960d provide four substreams of preconditioned symbols for four transmit antennas to buffers/multiplexers 970a through 970d, respectively.

[00243] Each buffer/multiplexer 970 receives pilot symbols and the preconditioned symbols from TX subband spatial processors 940a through 940k for the 48 data subbands. Each buffer/multiplexer 970 then, for each symbol period, multiplexes 4 pilot symbols, 48 preconditioned symbols, and 12 zeros for 4 pilot subbands, 48 data subbands, and 12 unused subbands, respectively, to form a sequence of 64 transmit symbols for that symbol period. Each buffer/multiplexer 970 provides a stream of transmit symbols $x_i(n)$ for one transmit antenna, where the transmit symbol stream comprises concatenated sequences of 64 transmit symbols. The transmit symbols can be scaled with correction factors to account for differences between the transmit/receive chains at the access point and the user terminal, as described below. The subsequent OFDM modulation for each transmit symbol stream is described above.

[00244] Parallel symbol streams may also be transmitted from the four transmit antennas without spatial processing at the access point using the non-steered spatial multiplexing mode. For this mode, the channel inversion process and beam-forming by beam-former 950 may be omitted. Each modulation symbol stream is further OFDM processed and transmitted from a respective transmit antenna.

[00245] The non-steered spatial multiplexing mode may be used for various situations such as if the transmitter is unable to perform the spatial processing necessary to support beam-steering based on eigenmode decomposition. This may be because the transmitter has not performed calibration procedures, is unable to generate a sufficiently good

estimate of the channel, or does not have calibration and eigenmode processing capabilities at all. For the non-steered spatial multiplexing mode, spatial multiplexing is still used to increase the transmission capacity, but the spatial processing to separate out the individual symbol streams is performed by the receiver.

[00246] For the non-steered spatial multiplexing mode, the receiver performs the spatial processing to recover the transmitted symbol streams. In particular, a user terminal may implement a channel correlation matrix inversion (CCMI) technique, a minimum mean square error (MMSE) technique, a successive interference cancellation receiver processing technique, or some other receiver spatial processing technique. These techniques are described in detail in commonly assigned U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001. The non-steered spatial multiplexing mode may be used for both downlink and uplink transmissions.

[00247] The multi-user spatial multiplexing mode supports data transmission to multiple user terminals simultaneously on the downlink based on the "spatial signatures" of the user terminals. The spatial signature for a user terminal is given by a channel response vector (for each subband) between the access point antennas and each user terminal antenna. The access point may obtain the spatial signatures, for example, based on the steered reference transmitted by the user terminals. The access point may process the spatial signatures for user terminals desiring data transmission to (1) select a set of user terminals for simultaneous data transmission on the downlink and (2) derive steering vectors for each of the independent data streams to be transmitted to the selected user terminals.

[00248] The steering vectors for the multi-user spatial multiplexing mode may be derived in various manners. Two exemplary schemes are described below. For simplicity, the following description is for one subband and assumes that each user terminal is equipped with a single antenna.

[00249] In a first scheme, the access point obtains the steering vectors using channel inversion. The access point may select N_{ap} single-antenna user terminals for simultaneous transmission on the downlink. The access point obtains an $1 \times N_{ap}$ channel response row vector for each selected user terminal and forms an $N_{ap} \times N_{ap}$ channel response matrix $\underline{\mathbf{H}}_{mu}$ with the N_{ap} row vectors for the N_{ap} user terminals. The

access point then obtains a matrix $\underline{\mathbf{H}}_{steer}$ of N_{ap} steering vectors for the N_{ap} selected user terminals as $\underline{\mathbf{H}}_{steer} = \underline{\mathbf{H}}_{mu}^{-1}$. The access point can also transmit a steered reference to each selected user terminal. Each user terminal processes its steered reference to estimate the channel gain and phase and coherently demodulates received symbols for its single antenna with the channel gain and phase estimates to obtain recovered symbols.

[00250] In a second scheme, the access point precodes N_{ap} symbol streams to be sent to N_{ap} user terminals such that these symbol streams experience little cross-talk at the user terminals. The access point can form the channel response matrix $\underline{\mathbf{H}}_{mu}$ for the N_{ap} selected user terminals and perform QR factorization on $\underline{\mathbf{H}}_{mu}$ such that $\underline{\mathbf{H}}_{mu} = \underline{\mathbf{F}}_{tri}\underline{\mathbf{Q}}_{mu}$, where $\underline{\mathbf{T}}_{tri}$ is a lower left triangular matrix $\underline{\mathbf{T}}_{tri}$ and $\underline{\mathbf{Q}}_{mu}$ is a unitary matrix. The access point then precodes the N_{ap} data symbol streams with the matrix $\underline{\mathbf{T}}_{tri}$ to obtain N_{ap} precoded symbol streams $\underline{\mathbf{a}}$, and further processes the precoded symbol streams with the unitary matrix $\underline{\mathbf{Q}}_{mu}$ to obtain the N_{ap} transmit symbol streams for transmission to the N_{ap} user terminals. Again, the access point can also transmit a steered reference to each user terminal. Each user terminal uses the steered reference to coherently demodulate its received symbols to obtain recovered symbols.

[00251] For the uplink in the multi-user spatial multiplexing mode, the access point can recover N_{ap} symbol streams transmitted simultaneously by N_{ap} user terminals using MMSE receiver processing, successive interference cancellation, or some other receiver processing technique. The access point can estimate the uplink channel response for each user terminal and use the channel response estimate for receiver spatial processing and for scheduling uplink transmissions. Each single-antenna user terminal can transmit an orthogonal pilot on the uplink. The uplink pilots from the N_{ap} user terminals can be orthogonal in time and/or frequency. Time orthogonality can be achieved by having each user terminal covers its uplink pilot with an orthogonal sequence assigned to the user terminal. Frequency orthogonality can be achieved by having each user terminal transmits its uplink pilot on a different set of subbands. The uplink transmissions from the user terminals should be approximately time-aligned at the access point (e.g., time-aligned to within the cyclic prefix).

3. Beam-Steering Mode - Transmit Processing

[00252] **FIG. 10A** shows a block diagram of a transmitter unit 1000 capable of performing the transmit processing for the beam-steering mode. Transmitter unit 1000 is yet another embodiment of the transmitter portion of the access point and the user terminal.

[00253] Within a TX data processor 710c, framing unit 808 frames the data for each FCH/RCH packet to generate one or more PHY frames for the packet. Scrambler 810 then scrambles the data for each transport channel. Encoder 812 next codes the framed data in accordance with a selected coding scheme to provide code bits. Puncture unit 814 then punctures the code bits to obtain the desired code rate for the wideband eigenmode used for data transmission. The code bits from puncture unit 818 are interleaved across all data subbands. Symbol mapping unit 820 then maps the interleaved data in accordance with a selected modulation scheme to provide modulation symbols. A TX spatial processor 720c then performs transmit processing on the modulation symbols for the beam-steering mode.

[00254] The beam-steering mode may be used to transmit data on one spatial channel or wideband eigenmode - typically the one associated with the largest eigenvalues for all of the data subbands. The beam-steering mode may be selected if the transmit power allocation to the wideband eigenmodes results in only one entry in the set $\underline{\boldsymbol{\alpha}}$ being non-zero. Whereas the spatial multiplexing mode performs beam-forming for each of the selected eigenmodes of each subband based on its eigenvector, the beam-steering mode performs beam-steering based on a "normalized" eigenvector for the principal eigenmode of each subband to transmit data on that single eigenmode.

[00255] The four elements of each eigenvector $\underline{\mathbf{v}}_1(k)$, for $k \in K$, for the principal eigenmode may have different magnitudes. The four preconditioned symbols obtained based on the four elements of eigenvector $\underline{\mathbf{v}}_1(k)$ for each subband may then have different magnitudes. Consequently, the four per-antenna transmit vectors, each of which includes the preconditioned symbols for all data subbands for a given transmit antenna, may have different magnitudes. If the transmit power for each transmit antenna is limited (e.g., because of limitations of the power amplifiers), then the beam-forming technique may not fully use the total power available for each antenna.

[00256] The beam-steering mode uses only the phase information from eigenvectors $\underline{\mathbf{v}}_1(k)$, for $k \in K$, for the principal eigenmode and normalizes each eigenvector such that all four elements in the eigenvector have equal magnitudes. The normalized eigenvector $\underline{\tilde{\mathbf{v}}}(k)$ for subband k may be expressed as:

$$\underline{\tilde{\mathbf{v}}}(k) = [Ae^{j\theta_1(k)} \quad Ae^{j\theta_2(k)} \quad Ae^{j\theta_3(k)} \quad Ae^{j\theta_4(k)}]^T \quad , \qquad \text{Eq (16)}$$

where A is a constant (e.g., $A=1$); and

$\theta_i(k)$ is the phase for subband k of transmit antenna i, which is given as:

$$\theta_i(k) = \angle v_{1,i}(k) = \tan^{-1}\left(\frac{\operatorname{Im}\{v_{1,i}(k)\}}{\operatorname{Re}\{v_{1,i}(k)\}}\right) . \qquad \text{Eq (17)}$$

As shown in equation (17), the phase of each element in the vector $\underline{\tilde{\mathbf{v}}}(k)$ is obtained from the corresponding element of eigenvector $\underline{\mathbf{v}}_1(k)$ (i.e., $\theta_i(k)$ is obtained from $v_{1,i}(k)$, where $\underline{\mathbf{v}}_1(k) = [v_{1,1}(k) \; v_{1,2}(k) \; v_{1,3}(k) \; v_{1,4}(k)]^T$).

[00257] Channel inversion may also be performed for the beam-steering mode so that a common rate can be used for all data subbands. The transmit power $\tilde{P}_1(k)$ allocated to each data subband for the beam-steering mode may be expressed as:

$$\tilde{P}_1(k) = \frac{\tilde{\beta}_1 \tilde{P}_1}{\tilde{\lambda}_1(k)} , \text{ for } k \in K , \qquad \text{Eq (18)}$$

where $\tilde{\beta}_1$ is a normalization factor that keeps the total transmit power invariant after channel inversion is applied;

$\tilde{P}_1$ is the transmit power allocated to each of the four antennas; and

$\tilde{\lambda}_1(k)$ is the power gain for subband k of the principal eigenmode for the beam-steering mode.

The normalization factor $\tilde{\beta}_1$ may be expressed as:

$$\tilde{\beta}_1 = \frac{1}{\sum_{k \in K} \tilde{\lambda}_1^{-1}(k)} . \qquad \text{Eq (19)}$$

The transmit power $\tilde{P}_1$ may be given as $P_1 = P_{total}/4$ (i.e., uniform allocation of the total transmit power across the four transmit antennas). The power gain $\tilde{\lambda}_1(k)$ may be expressed as:

$$\tilde{\lambda}_1(k) = \underline{\tilde{\mathbf{v}}}^H(k) \underline{\mathbf{H}}^H(k) \underline{\mathbf{H}}(k) \underline{\tilde{\mathbf{v}}}(k) . \qquad \text{Eq (20)}$$

[00258] The channel inversion results in power allocation of $\tilde{P}_1(k)$, for $k \in K$, for the 48 data subbands. The gain for each data subband may then be given as $\tilde{g}(k) = \sqrt{\tilde{P}_1(k)}$.

[00259] For the beam-steering mode, the transmit vector $\underline{\mathbf{x}}(k)$ for each subband may be expressed as:

$$\underline{\mathbf{x}}(k) = \underline{\tilde{\mathbf{v}}}(k)\tilde{g}(k)s(k) \quad \text{, for } k \in K . \qquad \text{Eq (21)}$$

Again for simplicity, equation (21) does not include the correction factors used to account for differences between the transmit/receive chains at the access point and the user terminal.

[00260] As shown in equation (16), the four elements of the normalized steering vector $\underline{\tilde{\mathbf{v}}}(k)$ for each subband have equal magnitude but possibly different phases. The beam-steering thus generates one transmit vector $\underline{\mathbf{x}}(k)$ for each subband, with the four elements of $\underline{\mathbf{x}}(k)$ having the same magnitude but possibly different phases.

[00261] **FIG. 10B** shows a block diagram of an embodiment of TX spatial processor 720c capable of performing the spatial processing for the beam-steering mode.

[00262] Within processor 720c, a demultiplexer 1032 receives and demultiplexes the modulation symbol stream $s(n)$ into 48 substreams for the 48 data subbands (denoted as $s(1)$ through $s(k)$). Each symbol substream is provided to a respective TX subband beam-steering processor 1040. Each processor 1040 performs the processing shown in equation (14) for one subband.

[00263] Within each TX subband beam-steering processor 1040, the modulation symbol substream is provided to a multiplier 1042, which also receives the gain $\tilde{g}(k)$ for the associated subband. Multiplier 1042 then scales the modulation symbols with the gain $\tilde{g}(k)$ to obtain scaled modulation symbols, which are then provided to a beam-steering unit 1050.

[00264] Beam-steering unit 1050 also receives the normalized eigenvector $\underline{\tilde{\mathbf{v}}}(k)$ for the associated subband. Within beam-steering unit 1050, the scaled modulation symbols are provided to four multipliers 1052a through 1052d, which also respectively receive the four elements, $\tilde{v}_1(k)$, $\tilde{v}_2(k)$, $\tilde{v}_3(k)$, and $\tilde{v}_4(k)$, of the normalized eigenvector $\underline{\tilde{\mathbf{v}}}(k)$. Each multiplier 1052 multiplies its scaled modulation symbols with its normalized eigenvector value $\tilde{v}_i(k)$ to provide preconditioned symbols. Multipliers 1052a through

1052d provide four preconditioned symbol substreams to buffers/multiplexers 1070a through 1070d, respectively.

[00265] Each buffer/multiplexer 1070 receives pilot symbols and the preconditioned symbols from TX subband beam-steering processors 1040a through 1040k for the 48 data subbands, multiplexes the pilot and preconditioned symbols and zeros for each symbol period, and provides a stream of transmit symbols $x_i(n)$ for one transmit antenna. The subsequent OFDM modulation for each transmit symbol stream is described above.

[00266] The processing for the beam-steering mode is described in further detail in commonly assigned U.S. Patent Application Serial No. 10/228,393, entitled "Beam-Steering and Beam-Forming for Wideband MIMO Systems," filed August 27, 2002. The system may also be designed to support a beam-forming mode whereby a data stream is transmitted on the principal eigenmode using the eigenvector instead of the normalized eigenvector.

4. <u>Framing for PHY frames</u>

[00267] **FIG. 11A** shows an embodiment of framing unit 808, which is used to frame the data for each FCH/RCH packet prior to subsequent processing by the TX data processor. This framing function may be bypassed for messages sent on the BCH, FCCH, and RACH. The framing unit generates an integer number of PHY frames for each FCH/RCH packet, where each PHY frame spans 6 OFDM symbols for the embodiment described herein.

[00268] For the diversity and beam-steering modes, only one spatial channel or wideband eigenmode is used for data transmission. The rate for this mode is known, and the number of information bits that may be sent in the payload of each PHY frame may be computed. For the spatial multiplexing mode, multiple spatial channels may be used for data transmission. Since the rate of each spatial channel is known, the number of information bits that may be sent in the payload of each PHY frame for all spatial channels may be computed.

[00269] As shown in FIG. 11A, the information bits (denoted as $i_1\, i_2\, i_3\, i_4 \ldots$) for each FCH/RCH packet are provided to a CRC generator 1102 and a multiplexer 1104 within framing unit 808. CRC generator 1102 generates a CRC value for the bits in the header (if any) and payload fields of each PHY frame and provides CRC bits to multiplexer

1104. Multiplexer 1104 receives the information bits, CRC bits, header bits, and pad bits (e.g., zeros), and provides these bits in the proper order, as shown in FIG. 6, based on a PHY Frame Control signal. The framing function may be bypassed by providing the information bits directly through multiplexer 1104. The framed or unframed bits (denoted as $d_1\, d_2\, d_3\, d_4 \ldots$) are provided to scrambler 810.

5. Scrambling

[00270] In an embodiment, the data bits for each transport channel are scrambled prior to coding. The scrambling randomizes the data so that a long sequence of all ones or all zeros is not transmitted. This can reduce the variation in the peak to average power of the OFDM waveform. The scrambling may be omitted for one or more transport channels and may also be selectively enabled and disabled.

[00271] FIG. 11A also shows an embodiment of scrambler 810. In this embodiment, scrambler 810 implements a generator polynomial:

$$G(x) = x^7 + x^4 + x \quad . \qquad \text{Eq (22)}$$

Other generator polynomials may also be used, and this is within the scope of the invention.

[00272] As shown in FIG. 11A, scrambler 810 includes seven delay elements 1112a through 1112g coupled in series. For each clock cycle, an adder 1114 performs modulo-2 addition of two bits stored in delay elements 1112d and 1112g and provides a scrambling bit to delay element 1112a.

[00273] The framed/unframed bits ($d_1\, d_2\, d_3\, d_4 \ldots$) are provided to an adder 1116, which also receives scrambling bits from adder 1114. Adder 1116 performs modulo-2 addition of each bit d_n with a corresponding scrambling bit to provide a scrambled bit q_n. Scrambler 810 provides a sequence of scrambled bits, which is denoted as $q_1\, q_2\, q_3\, q_4\, \ldots$.

[00274] The initial state of the scrambler (i.e., the content of delay elements 1112a through 1112g) is set to a 7-bit non-zero number at the start of each TDD frame. The three most significant bits (MSBs) (i.e., delay element 1112e through 1112f) are always set to one ('1') and the four least significant bits (LSBs) are set to the TDD frame counter, as indicated in the BCH message.

6. Encoding/Puncturing

[00275] In an embodiment, a single base code is used to code data prior to transmission. This base code generates code bits for one code rate. All other code rates supported by the system (as listed in Table 25) may be obtained by either repeating or puncturing the code bits.

[00276] **FIG. 11B** shows an embodiment of encoder 812 that implements the base code for the system. In this embodiment, the base code is a rate 1/2, constraint length 7 ($K = 7$), convolutional code with generators of 133 and 171 (octal).

[00277] Within encoder 812, a multiplexer 1120 receives and multiplexes the scrambled bits and tail bits (e.g., zeros). Encoder 812 further includes six delay elements 1122a through 1122f coupled in series. Four adders 1124a through 1124d are also coupled in series and used to implement the first generator (133). Similarly, four adders 1126a through 1126d are coupled in series and used to implement the second generator (171). The adders are further coupled to the delay elements in a manner to implement the two generators of 133 and 171, as shown in FIG. 11B.

[00278] The scrambled bits are provided to the first delay element 1122a and to adders 1124a and 1126a. For each clock cycle, adders 1124a through 1124d perform modulo-2 addition of the incoming bit and four prior bits stored in delay elements 1122b, 1122c, 1122e, and 1122f to provide the first code bit for that clock cycle. Similarly, adders 1126a through 1126d perform modulo-2 addition of the incoming bit and four prior bits stored in delay elements 1122a, 1122b, 1122c, and 1122f to provide the second code bit for that clock cycle. The code bits generated by the first generator are denoted as $a_1\, a_2\, a_3\, a_4 \ldots$, and the code bits generated by the second generator are denoted as $b_1\, b_2\, b_3\, b_4 \ldots$. A multiplexer 1128 then receives and multiplexes the two streams of code bits from the two generators into a single stream of code bits, which is denoted as $a_1\, b_1\, a_2\, b_2\, a_3\, b_3\, a_4\, b_4 \ldots$. For each scrambled bit q_n, two code bits a_n and b_n are generated, which results in a code rate of 1/2.

[00279] FIG. 11B also shows an embodiment of repeat/puncture unit 814 that can be used to generate other code rates based on the base code rate of 1/2. Within unit 814, the rate 1/2 code bits from encoder 812 are provided to a repeating unit 1132 and a puncturing unit 1134. Repeating unit 1132 repeats each rate 1/2 code bit once to obtain

an effective code rate of 1/4. Puncturing unit 1134 deletes some of the rate 1/2 code bits based on a specific puncturing pattern to provide the desired code rate.

[00280] Table 30 lists exemplary puncturing patterns that may be used for the various code rates supported by the system. Other puncturing patterns may also be used, and this is within the scope of the invention.

Table 30

Code Rate	Puncturing Pattern
1/2	11
7/12	11111110111110
5/8	1110111011
2/3	1110
11/16	1111101111111010011100
3/4	111001
13/16	01111011111101110000101100
5/6	1110011001
7/8	11101010011001

[00281] To obtain a code rate of k/n, puncturing unit 1134 provides n code bits for each group of $2k$ rate 1/2 code bits received from encoder 812. Thus, $2k-n$ code bits are deleted from each group of $2k$ code bits. The bits to be deleted from each group are denoted by zeros in the puncturing pattern. For example, to obtain a code rate of 7/12, two bits are deleted from each group of 14 code bits from encoder 812, with the deleted bits being the 8th and 14th code bits in the group, as denoted by the puncturing pattern of "11111110111110". No puncturing is performed if the desired code rate is 1/2.

[00282] A multiplexer 1136 receives the stream of code bits from repeating unit 1132 and the stream of code bits from puncturing unit 1134. Multiplexer 1136 then provides the code bits from repeating unit 1132 if the desired code rate is 1/4 and the code bits from puncturing unit 1134 if the desired code rate is 1/2 or higher.

[00283] Other codes and puncturing patterns besides those described above may also be used, and this is within the scope of the invention. For example, a Turbo code, a block code, some other codes, or any combination thereof may be used to code data. Also, different coding schemes may be used for different transport channels. For example,

convolutional coding may be used for the common transport channels, and Turbo coding may be used for the dedicated transport channels.

7. Interleaving

[00284] In an embodiment, the code bits to be transmitted are interleaved across the 48 data subbands. For the diversity and beam-steering modes, one stream of code bits is transmitted and interleaved across all data subbands. For the spatial multiplexing mode, up to four streams of code bits may be transmitted on up to four spatial channels. The interleaving may be performed separately for each spatial channel such that each stream of code bits is interleaved across all data subbands of the spatial channel used to transmit that stream. Table 29 shows an exemplary code bit-subband assignment that may be used for the interleaving for all transmission modes.

[00285] In one embodiment, the interleaving is performed across all 48 data subbands in each interleaving interval. For this embodiment, each group of 48 code bits in a stream is spread over the 48 data subbands to provide frequency diversity. The 48 code bits in each group may be assigned indices of 0 through 47. Each code bit index is associated with a respective subband. All code bits with a particular index are transmitted on the associated subband. For example, the first code bit (with index 0) in each group is transmitted on subband -26, the second code bit (with index 1) is transmitted on subband 1, the third code bit (with index 2) is transmitted on subband -17, and so on. This interleaving scheme may be used for the diversity, beam-steering, and spatial multiplexing modes. An alternative interleaving scheme for the spatial multiplexing mode is described below.

[00286] The interleaving may alternatively or additionally be performed over time. For example, after the interleaving across the data subbands, the code bits for each subband may further be interleaved (e.g., over one PHY frame or one PDU) to provide time diversity. For the spatial multiplexing mode, the interleaving may also be performed over multiple spatial channels.

[00287] Additionally, interleaving may be employed across the dimensions of the QAM symbols such that code bits forming QAM symbols are mapped to different bit positions of the QAM symbols.

8. Symbol Mapping

[00288] Table 31 shows the symbol mapping for various modulation schemes supported by the system. For each modulation scheme (except for BPSK), half of the bits are

mapped to an inphase (I) component and the other half of the bits are mapped to a quadrature (Q) component.

[00289] In an embodiment, the signal constellation for each supported modulation scheme may be defined based on Gray mapping. With Gray mapping, neighboring points in the signal constellation (in both the I and Q components) differ by only one bit position. Gray mapping reduces the number of bit errors for more likely error events, which corresponds to a received symbol being mapped to a location near the correct location, in which case only one code bit would be received in error.

Table 31

BPSK		
b	I	Q
0	-1	0
1	1	0

QPSK			
b_0	I	b_1	Q
0	-1	0	-1
1	1	1	1

16 QAM			
b_0b_1	I	b_2b_3	Q
00	-3	00	-3
01	-1	01	-1
11	1	11	1
10	3	10	3

64 QAM			
$b_0b_1b_2$	I	$b_3b_4b_5$	Q
000	-7	000	-7
001	-5	001	-5
011	-3	011	-3

256 QAM			
$b_0b_1b_2b_3$	I	$b_4b_5b_6b_7$	Q
0000	-15	0000	-15
0001	-13	0001	-13
0011	-11	0011	-11
0010	-9	0010	-9
0110	-7	0110	-7
0111	-5	0111	-5
0101	-3	0101	-3
0100	-1	0100	-1
1100	1	1100	1
1101	3	1101	3
1111	5	1111	5
1110	7	1110	7
1010	9	1010	9
1011	11	1011	11
1001	13	1001	13
1000	15	1000	15

Normalization Factor K_{norm}	
Modulation Scheme	Value

010	-1	010	-1
110	1	110	1
111	3	111	3
101	5	101	5
100	7	100	7

BPSK	1.0
QPSK	$1/\sqrt{2}$
16 QAM	$1/\sqrt{10}$
64 QAM	$1/\sqrt{42}$
256 QAM	$1/\sqrt{170}$

[00290] The I and Q values for each modulation scheme shown in Table 31 are scaled by a normalization factor K_{norm} so that the average power of all signal points in the associated signal constellation is equal to unity. The normalization factor for each modulation scheme is shown in Table 31. Quantized values for the normalization factors for the supported modulation schemes may also be used. A modulation symbol s from a particular signal constellation would then have the following form:

$$s = (\mathrm{I} + j\mathrm{Q}) \cdot K_{norm} \quad ,$$

where I and Q are the values in Table 31 for the signal constellation.

[00291] For a given PDU, the modulation may be different across the PDU and may be different for multiple spatial channels used for data transmission. For example, for the BCH PDU, different modulation schemes may be used for the beacon pilot, the MIMO pilot, and the BCH message.

9. <u>Processing for Spatial Multiplexing Mode</u>

[00292] For the spatial multiplexing mode, a PDU may be transmitted over multiple spatial channels. Various schemes may be used to process data for transmission over multiple spatial channels. Two specific processing schemes for the spatial multiplexing mode are described below.

[00293] In the first processing scheme, coding and puncturing are performed on a per spatial channel basis to achieve the desired code rate for each spatial channel. The N_E spatial channels to use for data transmission are ranked from the highest to lowest received SNR. The data for the entire PDU is first coded to obtain a stream of rate 1/2 code bits. The code bits are then punctured to obtain the desired code rate for each spatial channel.

[00294] The puncturing may be performed in sequential order for the N_E spatial channels, from the best (i.e., highest SNR) to the worst (i.e., lowest SNR) spatial

channel. In particular, the puncture unit first performs puncturing for the best spatial channel with the highest received SNR. When the correct number of code bits have been generated for the best spatial channel, the puncture unit then performs puncturing for the second best spatial channel with the next highest received SNR. This process continues until the code bits for all N_E spatial channels are generated. The order for puncturing is from the largest to smallest received SNR, regardless of the specific code rate used for each spatial channel.

[00295] For the example shown in Table 28, the 3456 information bits to be transmitted in the overall PHY frame are first coded with the rate 1/2 base code to obtain 6912 code bits. The first 3168 code bits are punctured using the puncturing pattern for code rate 11/16 to obtain 2304 code bits, which are provided in the PHY frame for the first spatial channel. The next 2592 code bits are then punctured using the puncturing pattern for code rate 3/4 to obtain 1728 code bits, which are provided in the PHY frame for the second spatial channel. The next 864 code bits are then punctured using the puncturing pattern for code rate 3/4 to obtain 576 code bits, which are provided in the PHY frame for the third spatial channel. The last 288 code bits for the PHY frame are then punctured using the puncturing pattern for code rate 1/2 to obtain 288 code bits, which are provided in the PHY frame for the last spatial channel. These four individual PHY frames are further processed and transmitted on the four spatial channels. The puncturing for the next overall PHY frame is then performed in similar manner. The first processing scheme may be implemented by TX data processor 710b in FIG. 9A.

[00296] In the second processing scheme, the coding and puncturing are performed for pairs of subbands. Moreover, the coding and puncturing are cycled through all selected spatial channels for each pair of subbands.

[00297] **FIG. 11C** shows a block diagram that illustrates a TX data processor 710d that implements the second processing scheme. Encoder 812 performs rate 1/2 convolutional encoding of the scrambled bits from scrambler 810. Each spatial channel is assigned a particular rate, which is associated with a specific combination of code rate and modulation scheme, as shown in Table 25. Let b_m denote the number of code bits per modulation symbol for spatial channel m (or equivalently, the number of code bits sent on each data subband of spatial channel m) and r_m denote the code rate used for spatial channel m. The value for b_m is dependent on the constellation size of the

modulation scheme used for spatial channel m. In particular, $b_m = 1, 2, 4, 6$ and 8 for BPSK, QPSK, 16-QAM, 64-QAM and 256-QAM, respectively.

[00298] Encoder 812 provides a stream of rate 1/2 code bits to demultiplexer 816, which demultiplexes the received code bit stream into four substreams for the four spatial channels. The demultiplexing is such that the first $4b_1r_1$ code bits are sent to buffer 813a for spatial channel 1, the next $4b_2r_2$ code bits are sent to buffer 813b for spatial channel 2, and so on. Each buffer 813 receives $4b_mr_m$ code bits each time demultiplexer 816 cycles through all four spatial channels. A total of $b_{total} = \sum_{m=1}^{4} 4b_mr_m$ rate 1/2 code bits are provided to the four buffers 813a through 813d for each cycle. Demultiplexer 816 thus cycles through all four positions for the four spatial channels for every b_{total} code bits, which is the number of code bits that can be sent on a pair of subbands using all four spatial channels.

[00299] Once each buffer 813 has been filled with $4b_mr_m$ code chips for the associated spatial channel, the code bits in the buffer are punctured to obtain the code rate for that spatial channel. Since $4b_mr_m$ rate 1/2 code bits span an integer number of puncturing periods for each puncturing pattern, exactly $2b_m$ code bits are provided after the puncturing for each spatial channel m. The $2b_m$ code bits for each spatial channel are then distributed (interleaved) over the data subbands.

[00300] In an embodiment, the interleaving is performed for each spatial channel in groups of 6 subbands at a time. The code bits after the puncturing for each spatial channel may be numbered sequentially as c_i, for $i = 0, 1, 2, \ldots$. A counter C_m may be maintained for each spatial channel to count every group of $6b_m$ code bits provided by the puncturing unit for that spatial channel. For example, for QPSK with $b_m = 2$, the counter would be set to $C_m = 0$ for code bits c_0 through c_{11} provided by the puncturing unit, $C_m = 1$ after code bits c_{12} through c_{23}, and so on. The counter value C_m for spatial channel m may be expressed as:

$$C_m = \lfloor i/(6b_m) \rfloor \bmod 8 \ . \qquad \text{Eq (23)}$$

[00301] To determine the subband to which code bit c_i is assigned, the bit index for the code bit is first determined as follows:

$$\text{bit index} = (i \bmod 6) + 6 \cdot C_m \quad . \qquad \text{Eq (24)}$$

The bit index is then mapped to the corresponding subband using Table 29.

[00302] For the example above, the first group of 6 code bits c_0 through c_5 is associated with bit indices 0 through 5, respectively, the second group of 6 code bits c_6 through c_{11} is also associated with bit indices 0 through 5, respectively. Code bits c_0 and c_6 would be mapped to subband -26, code bits c_1 and c_7 would be mapped to subband 1, and so on, as shown in Table 29. The spatial processing may then commence for this first group of 6 subbands. The third group of 6 code bits c_{12} through c_{17} (with $C_m = 1$) is associated with bit indices 6 through 11, respectively, and the fourth group of 6 code bits c_{18} through c_{23} is also associated with bit indices 6 through 11, respectively. Code bits c_{12} and c_{18} would be mapped to subband -25, code bits c_{13} and c_{19} would be mapped to subband 2, and so on. The spatial processing may then commence for this next group of 6 subbands.

[00303] The number 6 in equation (24) comes from the fact that the interleaving is performed in groups of six subbands. The (mod 8) operation in equation (23) comes from the fact that there are eight interleaving groups for the 48 data subbands. Since each cycle of demultiplexer 816 shown in FIG. 11C produces enough code bits to fill two subbands for each wideband eigenmode, a total of 24 cycles are needed to provide the $48b_m$ code bits for one OFDM symbol for each spatial channel.

[00304] The interleaving in groups of 6 subbands at a time can reduce processing delays. In particular, the spatial processing can commence once each group of 6 subbands is available.

[00305] In alternative embodiments, the interleaving may be performed for each spatial channel in groups of N_B subbands at a time, where N_B may be any integer (e.g., N_B may be equal to 48 for interleaving over all 48 data subbands).

VI. Calibration

[00306] For a TDD system, the downlink and uplink share the same frequency band in a time division duplexed manner. In this case, a high degree of correlation typically exists between the downlink and uplink channel responses. This correlation may be exploited to simplify the channel estimation and spatial processing. For a TDD system, each subband of the wireless link may be assumed to be reciprocal. That is, if $\underline{\mathbf{H}}(k)$

represents the channel response matrix from antenna array A to antenna array B for subband k, then a reciprocal channel implies that the coupling from array B to array A is given by the transpose of $\underline{\mathbf{H}}(k)$, which is $\underline{\mathbf{H}}^T(k)$.

[00307] However, the responses (gain and phase) of the transmit and receive chains at the access point are typically different from the responses of the transmit and receive chains at the user terminal. Calibration may be performed to determine the difference in the frequency responses of the transmit/receive chains at the access point and user terminal, and to account for the difference, such that the calibrated downlink and uplink responses can be expressed in terms of each other. Once the transmit/receive chains have been calibrated and accounted for, a measurement for one link (e.g., the downlink) may be used to derive steering vectors for the other link (e.g., the uplink).

[00308] The "effective" downlink and uplink channel responses, $\underline{\mathbf{H}}_{\mathrm{dn}}(k)$ and $\underline{\mathbf{H}}_{\mathrm{up}}(k)$, which include the responses of the applicable transmit and receive chains at the access point and user terminal, may be expressed as:

$$\underline{\mathbf{H}}_{\mathrm{dn}}(k) = \underline{\mathbf{R}}_{\mathrm{ut}}(k)\underline{\mathbf{H}}(k)\underline{\mathbf{T}}_{\mathrm{ap}}(k) \ , \text{ for } k \in K \text{, and} \qquad \text{Eq (25)}$$

$$\underline{\mathbf{H}}_{\mathrm{up}}(k) = \underline{\mathbf{R}}_{\mathrm{ap}}(k)\underline{\mathbf{H}}^T(k)\underline{\mathbf{T}}_{\mathrm{ut}}(k) \ , \text{ for } k \in K,$$

where $\underline{\mathbf{T}}_{\mathrm{ap}}(k)$ and $\underline{\mathbf{R}}_{\mathrm{ap}}(k)$ are $N_{ap} \times N_{ap}$ diagonal matrices with entries for the complex gains associated with the transmit chain and receive chain, respectively, for the N_{ap} antennas at the access point for subband k;

$\underline{\mathbf{T}}_{\mathrm{ut}}(k)$ and $\underline{\mathbf{R}}_{\mathrm{ut}}(k)$ are $N_{ut} \times N_{ut}$ diagonal matrices with entries for the complex gains associated with the transmit chain and receive chain, respectively, for the N_{ut} antennas at the user terminal for subband k; and

$\underline{\mathbf{H}}(k)$ is an $N_{ut} \times N_{ap}$ channel response matrix for the downlink.

[00309] Combining the two equations in equation set (25), the following relationship may be obtained:

$$\underline{\mathbf{H}}_{\mathrm{up}}(k)\underline{\mathbf{K}}_{\mathrm{ut}}(k) = (\underline{\mathbf{H}}_{\mathrm{dn}}(k)\underline{\mathbf{K}}_{\mathrm{ap}}(k))^T \ , \text{ for } k \in K, \qquad \text{Eq (26)}$$

where $\underline{\mathbf{K}}_{\mathrm{ut}}(k) = \underline{\mathbf{T}}_{\mathrm{ut}}^{-1}(k)\underline{\mathbf{R}}_{\mathrm{ut}}(k)$ and $\underline{\mathbf{K}}_{\mathrm{ap}}(k) = \underline{\mathbf{T}}_{\mathrm{ap}}^{-1}(k)\underline{\mathbf{R}}_{\mathrm{ap}}(k)$.

[00310] The left-hand side of equation (26) represents the "true" calibrated channel response on the uplink, and the right-hand side represents the transpose of the "true" calibrated channel response on the downlink. The application of the diagonal matrices

$\underline{\mathbf{K}}_{ap}(k)$ and $\underline{\mathbf{K}}_{ut}(k)$ to the effective downlink and uplink channel responses, respectively, as shown in equation (26), allows the calibrated channel responses for the downlink and uplink to be expressed as transposes of each other. The $(N_{ap} \times N_{ap})$ diagonal matrix $\underline{\mathbf{K}}_{ap}(k)$ for the access point is the ratio of the receive chain response $\underline{\mathbf{R}}_{ap}(k)$ to the transmit chain response $\underline{\mathbf{T}}_{ap}(k)$ (i.e., $\underline{\mathbf{K}}_{ap}(k) = \frac{\underline{\mathbf{R}}_{ap}(k)}{\underline{\mathbf{T}}_{ap}(k)}$), where the ratio is taken element-by-element. Similarly, the $(N_{ut} \times N_{ut})$ diagonal matrix $\underline{\mathbf{K}}_{ut}(k)$ for the user terminal is the ratio of the receive chain response $\underline{\mathbf{R}}_{ut}(k)$ to the transmit chain response $\underline{\mathbf{T}}_{ut}(k)$.

[00311] The matrices $\underline{\mathbf{K}}_{ap}(k)$ and $\underline{\mathbf{K}}_{ut}(k)$ include values that can account for differences in the transmit/receive chains at the access point and the user terminal. This would then allow the channel response for one link to be expressed by the channel response for the other link, as shown in equation (26).

[00312] Calibration may be performed to determine the matrices $\underline{\mathbf{K}}_{ap}(k)$ and $\underline{\mathbf{K}}_{ut}(k)$. Typically, the true channel response $\underline{\mathbf{H}}(k)$ and the transmit/receive chain responses are not known nor can they be exactly or easily ascertained. Instead, the effective downlink and uplink channel responses, $\underline{\mathbf{H}}_{dn}(k)$ and $\underline{\mathbf{H}}_{up}(k)$, may be estimated based on pilots sent on the downlink and uplink, respectively, as described below. Estimates of the matrices $\underline{\mathbf{K}}_{ap}(k)$ and $\underline{\mathbf{K}}_{ut}(k)$, which are referred to as correction matrices $\hat{\underline{\mathbf{K}}}_{ap}(k)$ and $\hat{\underline{\mathbf{K}}}_{ut}(k)$, may then be derived based on the downlink and uplink channel response estimates, $\hat{\underline{\mathbf{H}}}_{dn}(k)$ and $\hat{\underline{\mathbf{H}}}_{up}(k)$, as described below. The matrices $\hat{\underline{\mathbf{K}}}_{ap}(k)$ and $\hat{\underline{\mathbf{K}}}_{ut}(k)$ include correction factors that can account for differences in the transmit/receive chains at the access point and the user terminal.

[00313] The "calibrated" downlink and uplink channel responses observed by the user terminal and the access point, respectively, may then be expressed as:

$$\underline{\mathbf{H}}_{cdn}(k) = \underline{\mathbf{H}}_{dn}(k)\hat{\underline{\mathbf{K}}}_{ap}(k) \quad \text{, for } k \in K \text{, and} \qquad \text{Eq (27)}$$

$$\underline{\mathbf{H}}_{cup}(k) = \underline{\mathbf{H}}_{up}(k)\hat{\underline{\mathbf{K}}}_{ut}(k) \quad \text{, for } k \in K \text{,}$$

where $\underline{\mathbf{H}}_{\text{cdn}}^{T}(k)$ and $\underline{\mathbf{H}}_{\text{cup}}(k)$ are estimates of the "true" calibrated channel response expressions in equation (26). Combining the two equations in equation set (27) using the expression in equation (26), it can be shown that $\underline{\mathbf{H}}_{\text{cup}}(k) \approx \underline{\mathbf{H}}_{\text{cdn}}^{T}(k)$. The accuracy of the relationship $\underline{\mathbf{H}}_{\text{cup}}(k) \approx \underline{\mathbf{H}}_{\text{cdn}}^{T}(k)$ is dependent on the accuracy of the matrices $\underline{\hat{\mathbf{K}}}_{\text{ap}}(k)$ and $\underline{\hat{\mathbf{K}}}_{\text{ut}}(k)$, which in turn is typically dependent on the quality of the downlink and uplink channel response estimates, $\underline{\hat{\mathbf{H}}}_{\text{dn}}(k)$ and $\underline{\hat{\mathbf{H}}}_{\text{up}}(k)$.

[00314] The calibration may be performed using various schemes. For clarity, a specific calibration scheme is described below. To perform the calibration, the user terminal initially acquires the timing and frequency of the access point based on the beacon pilot transmitted on the BCH. The user terminal then sends a message on the RACH to initiate a calibration procedure with the access point. The calibration may be performed in parallel with registration/ authentication.

[00315] Since the frequency responses of the transmit/receive chains at the access point and user terminal are typically flat over most of the band of interest, the phase/gain differences of the transmit/receive chains may be characterized with a small number of subbands. The calibration may be performed for 4, 8, 16, 48 or some other number of subbands, which may be specified in the message sent to initiate the calibration. Calibration may also be performed for the pilot subbands. Calibration constants for subbands on which calibration is not explicitly performed may be computed by interpolation on calibrated subbands. For clarity, the following assumes that calibration is performed for all data subbands.

[00316] For the calibration, the access point allocates to the user terminal a sufficient amount of time on the RCH to send an uplink MIMO pilot of sufficient duration plus a message. The duration of the uplink MIMO pilot may be dependent on the number of subbands over which calibration is performed. For example, 8 OFDM symbols may be sufficient if calibration is performed for four subbands, and more (e.g., 20) OFDM symbols may be needed for more subbands. The total transmit power is typically fixed, so if the MIMO pilot is transmitted on a small number of subbands, then higher amounts of transmit power may be used for each of these subbands and the SNR for each subband is high. Conversely, if the MIMO pilot is transmitted on a large number of subbands then smaller amounts of transmit power may be used for each subband and the

SNR for each subband is worse. If the SNR of each subband is not sufficiently high, then more OFDM symbols may be sent for the MIMO pilot and integrated at the receiver to obtain a higher overall SNR for the subband.

[00317] The user terminal then transmits a MIMO pilot on the RCH, which is used by the access point to derive an estimate of the effective uplink channel response, $\hat{\underline{\mathbf{H}}}_{\text{up}}(k)$, for each of the data subbands. The uplink channel response estimates are quantized (e.g., to 12-bit complex values, with inphase (I) and quadrature (Q) components) and sent to the user terminal.

[00318] The user terminal also derives an estimate of the effective downlink channel response, $\hat{\underline{\mathbf{H}}}_{\text{dn}}(k)$, for each of the data subbands based on the downlink MIMO pilot sent on the BCH. Upon obtaining the effective uplink and downlink channel response estimates, $\hat{\underline{\mathbf{H}}}_{\text{up}}(k)$ and $\hat{\underline{\mathbf{H}}}_{\text{dn}}(k)$, for all data subbands, the user terminal determines correction factors, $\hat{\underline{\mathbf{K}}}_{\text{ap}}(k)$ and $\hat{\underline{\mathbf{K}}}_{\text{ut}}(k)$, for each of the data subbands, which are to be used by the access point and user terminal, respectively. A correction vector $\hat{\underline{\mathbf{k}}}_{\text{ap}}(k)$ may be defined to include only the diagonal elements of $\hat{\underline{\mathbf{K}}}_{\text{ap}}(k)$, and a correction vector $\hat{\underline{\mathbf{k}}}_{\text{ut}}(k)$ may be defined to include only the diagonal elements of $\hat{\underline{\mathbf{K}}}_{\text{ut}}(k)$.

[00319] The correction factors may be derived in various manners, including by a matrix-ratio computation and an MMSE computation. Both of these computation methods are described in further detail below. Other computation methods may also be used, and this is within the scope of the invention.

1. <u>Matrix-Ratio Computation</u>

[00320] To determine the correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap}}(k)$ and $\hat{\underline{\mathbf{k}}}_{\text{ut}}(k)$ given the effective downlink and uplink channel response estimates, $\hat{\underline{\mathbf{H}}}_{\text{dn}}(k)$ and $\hat{\underline{\mathbf{H}}}_{\text{up}}(k)$, an $(N_{ut} \times N_{ap})$ matrix $\underline{\mathbf{C}}(k)$ is first computed for each data subband, as follows:

$$\underline{\mathbf{C}}(k) = \frac{\hat{\underline{\mathbf{H}}}_{\text{up}}^{T}(k)}{\hat{\underline{\mathbf{H}}}_{\text{dn}}(k)}, \text{ for } k \in K, \qquad \text{Eq (28)}$$

where the ratio is taken element-by-element. Each element of $\underline{\mathbf{C}}(k)$ may thus be computed as:

$$c_{i,j}(k) = \frac{\hat{h}_{\mathrm{up}\,i,j}(k)}{\hat{h}_{\mathrm{dn}\,i,j}(k)} \quad , \text{ for } i = \{1 \;...\; N_{ut}\} \text{ and } j = \{1 \;...\; N_{ap}\}, \qquad \text{Eq (29)}$$

where $\hat{h}_{\mathrm{up}\,i,j}(k)$ is the (i,j)-th (row, column) element of $\underline{\hat{\mathbf{H}}}_{\mathrm{up}}^{T}(k)$, $\hat{h}_{\mathrm{dn}\,i,j}(k)$ is the (i,j)-th element of $\underline{\hat{\mathbf{H}}}_{\mathrm{dn}}(k)$, and $c_{i,j}(k)$ is the (i,j)-th element of $\underline{\mathbf{C}}(k)$.

[00321] The correction vector $\underline{\hat{\mathbf{k}}}_{\mathrm{ap}}(k)$ for the access point is then equal to the mean of the normalized rows of $\underline{\mathbf{C}}(k)$. Each row of $\underline{\mathbf{C}}(k)$ is first normalized by scaling each of the N_{ap} elements in the row with the first element in the row. Thus, if $\underline{\mathbf{c}}_i(k) = [c_{i,1}(k) \;...\; c_{i,N_{ap}}(k)]$ is the i-th row of $\underline{\mathbf{C}}(k)$, then the normalized row $\underline{\tilde{\mathbf{c}}}_i(k)$ may be expressed as:

$$\underline{\tilde{\mathbf{c}}}_i(k) = [c_{i,1}(k)/c_{i,1}(k) \;...\; c_{i,j}(k)/c_{i,1}(k) \;...\; c_{i,N_{ap}}(k)/c_{i,1}(k)] \; . \qquad \text{Eq (30)}$$

The mean of the normalized rows is then the sum of the N_{ut} normalized rows divided by N_{ut}, which may be expressed as:

$$\underline{\hat{\mathbf{k}}}_{\mathrm{ap}}(k) = \frac{1}{N_{ut}} \sum_{i=1}^{N_{ut}} \underline{\tilde{\mathbf{c}}}_i(k) \quad , \text{for } k \in K \, . \qquad \text{Eq (31)}$$

Because of the normalization, the first element of $\underline{\hat{\mathbf{k}}}_{\mathrm{ap}}(k)$ is unity.

[00322] The correction vector $\underline{\hat{\mathbf{k}}}_{\mathrm{ut}}(k)$ for the user terminal is equal to the mean of the inverses of the normalized columns of $\underline{\mathbf{C}}(k)$. The j-th column of $\underline{\mathbf{C}}(k)$ is first normalized by scaling each element in the column with the j-th element of the vector $\underline{\hat{\mathbf{k}}}_{\mathrm{ap}}(k)$, which is denoted as $K_{\mathrm{ap},j,j}(k)$. Thus, if $\underline{\mathbf{c}}_j(k) = [c_{1,j}(k) \;...\; c_{N_{ut},j}(k)]^T$ is the j-th column of $\underline{\mathbf{C}}(k)$, then the normalized column $\underline{\breve{\mathbf{c}}}_j(k)$ may be expressed as:

$$\underline{\breve{\mathbf{c}}}_j(k) = [c_{1,j}(k)/K_{\mathrm{ap},j,j}(k) \;...\; c_{i,j}(k)/K_{\mathrm{ap},j,j}(k) \;...\; c_{N_{ut},j}(k)/K_{\mathrm{ap},j,j}(k)]^T \, . \qquad \text{Eq (32)}$$

The mean of the inverses of the normalized columns is then the sum of the inverses of the N_{ap} normalized columns divided by N_{ap}, which may be expressed as:

$$\underline{\hat{\mathbf{k}}}_{\mathrm{ut}}(k) = \frac{1}{N_{ap}} \sum_{j=1}^{N_{ap}} \frac{1}{\underline{\breve{\mathbf{c}}}_j(k)} \quad , \text{for } k \in K \, , \qquad \text{Eq (33)}$$

where the inversion of the normalized columns, $\underline{\breve{\mathbf{c}}}_j(k)$, is performed element-wise.

2. MMSE Computation

[00323] For the MMSE computation, the correction factors $\underline{\hat{\mathbf{K}}}_{\rm ap}(k)$ and $\underline{\hat{\mathbf{K}}}_{\rm ut}(k)$ are derived from the effective downlink and uplink channel response estimates, $\underline{\hat{\mathbf{H}}}_{\rm dn}(k)$ and $\underline{\hat{\mathbf{H}}}_{\rm up}(k)$, such that the mean square error (MSE) between the calibrated downlink channel response and the calibrated uplink channel response is minimized. This condition may be expressed as:

$$\min \left|(\underline{\hat{\mathbf{H}}}_{\rm dn}(k)\underline{\hat{\mathbf{K}}}_{\rm ap}(k))^T - (\underline{\hat{\mathbf{H}}}_{\rm up}(k)\underline{\hat{\mathbf{K}}}_{\rm ut}(k))\right|^2 \text{ , for } k \in K\,, \qquad \text{Eq (34)}$$

which may also be written as:

$$\min \left|\underline{\hat{\mathbf{K}}}_{\rm ap}(k)\underline{\hat{\mathbf{H}}}_{\rm dn}^T(k) - \underline{\hat{\mathbf{H}}}_{\rm up}(k)\underline{\hat{\mathbf{K}}}_{\rm ut}(k)\right|^2 \text{ , for } k \in K\,,$$

where $\underline{\hat{\mathbf{K}}}_{\rm ap}^T(k) = \underline{\hat{\mathbf{K}}}_{\rm ap}(k)$ since $\underline{\hat{\mathbf{K}}}_{\rm ap}(k)$ is a diagonal matrix.

[00324] Equation (34) is subject to the constraint that the lead element of $\underline{\hat{\mathbf{K}}}_{\rm ap}(k)$ is set equal to unity (i.e., $\underline{\hat{\mathbf{K}}}_{\rm ap,0,0}(k) = 1$). Without this constraint, the trivial solution would be obtained with all elements of the matrices $\underline{\hat{\mathbf{K}}}_{\rm ap}(k)$ and $\underline{\hat{\mathbf{K}}}_{\rm ut}(k)$ set equal to zero. In equation (34), a matrix $\underline{\mathbf{Y}}(k)$ is first obtained as $\underline{\mathbf{Y}}(k) = \underline{\hat{\mathbf{K}}}_{\rm ap}(k)\underline{\hat{\mathbf{H}}}_{\rm dn}^T(k) - \underline{\hat{\mathbf{H}}}_{\rm up}(k)\underline{\hat{\mathbf{K}}}_{\rm ut}(k)$. The square of the absolute value is next obtained for each of the $N_{ap} \cdot N_{ut}$ entries of the matrix $\underline{\mathbf{Y}}(k)$. The mean square error (or the square error, since a divide by $N_{ap} \cdot N_{ut}$ is omitted) is then equal to the sum of all $N_{ap} \cdot N_{ut}$ squared values.

[00325] The MMSE computation is performed for each designated subband to obtain the correction factors $\underline{\hat{\mathbf{K}}}_{\rm ap}(k)$ and $\underline{\hat{\mathbf{K}}}_{\rm ut}(k)$ for that subband. The MMSE computation for one subband is described below. For simplicity, the subband index, *k*, is omitted in the following description. Also for simplicity, the elements of the downlink channel response estimate $\underline{\hat{\mathbf{H}}}_{\rm dn}^T$ are denoted as $\{a_{ij}\}$, the elements of the uplink channel response estimate $\underline{\hat{\mathbf{H}}}_{\rm up}$ are denoted as $\{b_{ij}\}$, the diagonal elements of the matrix $\underline{\hat{\mathbf{K}}}_{\rm ap}$ are denoted as $\{u_i\}$, and the diagonal elements of the matrix $\underline{\hat{\mathbf{K}}}_{\rm ut}$ are denoted as $\{v_j\}$, where $i = \{1 \;...\; N_{ap}\}$ and $j = \{1 \;...\; N_{ut}\}$.

[00326] The mean square error may be rewritten from equation (34), as follows:

$$\mathrm{MSE} = \sum_{j=1}^{N_{ut}} \sum_{i=1}^{N_{ap}} | a_{ij} u_i - b_{ij} v_j |^2 \quad , \qquad \text{Eq (35)}$$

again subject to the constraint $u_1 = 1$. The minimum mean square error may be obtained by taking the partial derivatives of equation (35) with respect to u and v and setting the partial derivatives to zero. The results of these operations are the following equation sets:

$$\sum_{j=1}^{N_{ut}} (a_{ij} u_i - b_{ij} v_j) \cdot a_{ij}^* = 0 \quad \text{, for } i \in \{2 \ \dots \ N_{ap}\}\text{, and} \qquad \text{Eq (36a)}$$

$$\sum_{i=1}^{N_{ap}} (a_{ij} u_i - b_{ij} v_j) \cdot b_{ij}^* = 0 \quad \text{, for } j \in \{1 \ \dots \ N_{ut}\}. \qquad \text{Eq (36b)}$$

In equation (36a), $u_1 = 1$ so there is no partial derivative for this case, and the index i runs from 2 through N_{ap}.

[00327] The set of $(N_{ap} + N_{ut} - 1)$ equations in equation sets (36a) and (36b) may be more conveniently expressed in matrix form, as follows:

$$\underline{\mathbf{A}}\underline{\mathbf{y}} = \underline{\mathbf{z}} \quad , \qquad \text{Eq (37)}$$

where

$$\underline{\mathbf{A}} = \begin{bmatrix}
\sum_{j=1}^{N_{ut}} |a_{2j}|^2 & 0 & \dots & 0 & -b_{21}a_{21}^* & \dots & -b_{2N_{ap}}a_{2N_{ut}}^* \\
0 & \sum_{j=1}^{N_{ut}} |a_{3j}|^2 & 0 & \dots & \dots & \dots & \dots \\
\dots & 0 & \dots & 0 & & & \\
0 & \dots & 0 & \sum_{j=1}^{N_{ut}} |a_{N_{ap}j}|^2 & -b_{N_{ap}1}a_{N_{ap}1}^* & & -b_{N_{ap}N_{ut}}a_{N_{ap}N_{ut}}^* \\
-a_{21}b_{21}^* & \dots & & -a_{N_{ap}1}b_{N_{ap}1}^* & \sum_{i=1}^{N_{ap}} |b_{i1}|^2 & 0 \quad \dots & 0 \\
\dots & \dots & & & 0 & \sum_{i=1}^{N_{ap}} |b_{i2}|^2 \quad 0 & \dots \\
 & & & & \dots & 0 \quad \dots & 0 \\
-a_{2N_{ut}}b_{2N_{ut}}^* & \dots & & -a_{N_{ap}N_{ut}}b_{N_{ap}N_{ut}}^* & 0 & \dots \quad 0 & \sum_{i=1}^{N_{ap}} |b_{iN_{ut}}|^2
\end{bmatrix}$$

$$\underline{\mathbf{y}} = \begin{bmatrix} u_2 \\ u_3 \\ \cdots \\ u_{N_{ap}} \\ v_1 \\ v_2 \\ \cdots \\ v_{N_{ut}} \end{bmatrix} \text{ and } \underline{\mathbf{z}} = \begin{bmatrix} 0 \\ 0 \\ \cdots \\ 0 \\ a_{11}b_{11}^* \\ a_{12}b_{12}^* \\ \cdots \\ a_{1N_{ut}}b_{1N_{ut}}^* \end{bmatrix} .$$

[00328] The matrix $\underline{\mathbf{A}}$ includes $(N_{ap}+N_{ut}-1)$ rows, with the first $N_{ap}-1$ rows corresponding to the $N_{ap}-1$ equations from equation set (36a) and the last N_{ut} rows corresponding to the N_{ut} equations from equation set (36b). In particular, the first row of the matrix $\underline{\mathbf{A}}$ is generated fr0m equation set (36a) with $i=2$, the second row is generated with $i=3$, and so on. The N_{ap}-th row of the matrix $\underline{\mathbf{A}}$ is generated from equation set (36b) with $j=1$, and so on, and the last row is generated with $j=N_{ut}$. As shown above, the entries of the matrix $\underline{\mathbf{A}}$ and the entries of the vector $\underline{\mathbf{z}}$ may be obtained based on the entries in the matrices $\hat{\underline{\mathbf{H}}}_{\text{dn}}^T$ and $\hat{\underline{\mathbf{H}}}_{\text{up}}$.

[00329] The correction factors are included in the vector $\underline{\mathbf{y}}$, which may be obtained as:

$$\underline{\mathbf{y}} = \underline{\mathbf{A}}^{-1}\underline{\mathbf{z}} \ . \qquad \text{Eq (38)}$$

[00330] The results of the MMSE computation are correction matrices $\hat{\underline{\mathbf{K}}}_{\text{ap}}$ and $\hat{\underline{\mathbf{K}}}_{\text{ut}}$ that minimize the mean square error in the calibrated downlink and uplink channel responses, as shown in equation (34). Since the matrices $\hat{\underline{\mathbf{K}}}_{\text{ap}}$ and $\hat{\underline{\mathbf{K}}}_{\text{ut}}$ are obtained based on the downlink and uplink channel response estimates, $\hat{\underline{\mathbf{H}}}_{\text{dn}}$ and $\hat{\underline{\mathbf{H}}}_{\text{up}}$, the quality of the correction matrices $\hat{\underline{\mathbf{K}}}_{\text{ap}}$ and $\hat{\underline{\mathbf{K}}}_{\text{ut}}$ are thus dependent on the quality of the channel estimates $\hat{\underline{\mathbf{H}}}_{\text{dn}}$ and $\hat{\underline{\mathbf{H}}}_{\text{up}}$. The MIMO pilot may be averaged at the receiver to obtain more accurate estimates for $\hat{\underline{\mathbf{H}}}_{\text{dn}}$ and $\hat{\underline{\mathbf{H}}}_{\text{up}}$.

[00331] The correction matrices, $\hat{\underline{\mathbf{K}}}_{\text{ap}}$ and $\hat{\underline{\mathbf{K}}}_{\text{ut}}$, obtained based on the MMSE computation are generally better than the correction matrices obtained based on the

matrix-ratio computation. This is especially true when some of the channel gains are small and measurement noise can greatly degrade the channel gains.

3. Post Computation

[00332] A pair of correction vectors, $\hat{\underline{\mathbf{k}}}_{\text{ap}}(k)$ and $\hat{\underline{\mathbf{k}}}_{\text{ut}}(k)$, may be determined for each of the data subbands. Since adjacent subbands are likely to be correlated, the computation may be simplified. For example, the computation may be performed for every n-th subband instead of each subband, where n may be determined by the expected response of the transmit/receive chains. If the calibration is performed for fewer than all of the data and pilot subbands, then the correction factors for the "uncalibrated" subbands may be obtained by interpolating the correction factors obtained for the "calibrated" subbands.

[00333] Various other calibration schemes may also be used to derive the correction vectors, $\hat{\underline{\mathbf{k}}}_{\text{ap}}(k)$ and $\hat{\underline{\mathbf{k}}}_{\text{ut}}(k)$, for the access point and the user terminal, respectively. However, the scheme described above allows "compatible" correction vectors to be derived for the access point when the calibration is performed by different user terminals.

[00334] After the derivation, the user terminal sends the correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap}}(k)$ for all data subbands back to the access point. If the access point has already been calibrated (e.g., by other user terminals), then the current correction vectors are updated with the newly received correction vectors. Thus, if the access point uses correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap1}}(k)$ to transmit the MIMO pilot from which the user terminal determines new correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap2}}(k)$, then the updated correction vectors are the product of the current and new correction vectors, i.e., $\hat{\underline{\mathbf{k}}}_{\text{ap3}}(k) = \hat{\underline{\mathbf{k}}}_{\text{ap1}}(k) \cdot \hat{\underline{\mathbf{k}}}_{\text{ap2}}(k)$, where the multiplication is element-by-element. The updated correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap3}}(k)$ may then be used by the access point until they are updated again.

[00335] The correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap1}}(k)$ and $\hat{\underline{\mathbf{k}}}_{\text{ap2}}(k)$ may be derived by the same or different user terminals. In one embodiment, the updated correction vectors are defined as $\hat{\underline{\mathbf{k}}}_{\text{ap3}}(k) = \hat{\underline{\mathbf{k}}}_{\text{ap1}}(k) \cdot \hat{\underline{\mathbf{k}}}_{\text{ap2}}(k)$, where the multiplication is element-by-element. In another embodiment, the updated correction vectors may be redefined as

$\hat{\underline{\mathbf{k}}}_{\text{ap3}}(k) = \hat{\underline{\mathbf{k}}}_{\text{ap1}}(k) \cdot \hat{\underline{\mathbf{k}}}^{\alpha}_{\text{ap2}}(k)$, where α is a factor used to provide weighted averaging (e.g., $0 < \alpha < 1$). If the calibration updates are infrequent, then □ close to one might perform best. If the calibration updates are frequent but noisy, then a smaller value for □ is better. The updated correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap3}}(k)$ may then be used by the access point until they are updated again.

[00336] The access point and user terminal use their respective correction vectors $\hat{\underline{\mathbf{k}}}_{\text{ap}}(k)$ and $\hat{\underline{\mathbf{k}}}_{\text{ut}}(k)$, or the corresponding correction matrices $\hat{\underline{\mathbf{K}}}_{\text{ap}}(k)$ and $\hat{\underline{\mathbf{K}}}_{\text{ut}}(k)$, for $k \in K$, to scale the modulation symbols prior to transmission, as described below. The calibrated downlink and uplink channels that the user terminal and access point observe are shown in equation (27).

VII. Spatial Processing

[00337] The spatial processing at the access point and user terminal may be simplified for a TDD system, after calibration has been performed to account for the difference in the transmit/receive chains. As noted above, the calibrated downlink channel response is $\underline{\mathbf{H}}_{\text{cdn}}(k) = \underline{\mathbf{H}}_{\text{dn}}(k)\hat{\underline{\mathbf{K}}}_{\text{ap}}(k)$. The calibrated uplink channel response is $\underline{\mathbf{H}}_{\text{cup}}(k) = \underline{\mathbf{H}}_{\text{up}}(k)\hat{\underline{\mathbf{K}}}_{\text{ut}}(k) \approx (\underline{\mathbf{H}}_{\text{dn}}(k)\hat{\underline{\mathbf{K}}}_{\text{ap}}(k))^T$.

1. Uplink Spatial Processing

[00338] Singular value decomposition of the calibrated uplink channel response matrix, $\underline{\mathbf{H}}_{\text{cup}}(k)$, may be expressed as:

$$\underline{\mathbf{H}}_{\text{cup}}(k) = \underline{\mathbf{U}}_{\text{ap}}(k)\underline{\Sigma}(k)\underline{\mathbf{V}}^{H}_{\text{ut}}(k) \quad \text{, for } k \in K, \qquad \text{Eq (39)}$$

where $\underline{\mathbf{U}}_{\text{ap}}(k)$ is an $(N_{ap} \times N_{ap})$ unitary matrix of left eigenvectors of $\underline{\mathbf{H}}_{\text{cup}}(k)$;

$\underline{\Sigma}(k)$ is an $(N_{ap} \times N_{ut})$ diagonal matrix of singular values of $\underline{\mathbf{H}}_{\text{cup}}(k)$, and

$\underline{\mathbf{V}}_{\text{ut}}(k)$ is an $(N_{ut} \times N_{ut})$ unitary matrix of right eigenvectors of $\underline{\mathbf{H}}_{\text{cup}}(k)$.

[00339] Correspondingly, singular value decomposition of the calibrated downlink channel response matrix, $\underline{\mathbf{H}}_{\text{cdn}}(k)$, may be expressed as:

$$\underline{\mathbf{H}}_{\text{cdn}}(k) = \underline{\mathbf{V}}^{*}_{\text{ut}}(k)\underline{\Sigma}(k)\underline{\mathbf{U}}^{T}_{\text{ap}}(k) \quad \text{, for } k \in K. \qquad \text{Eq (40)}$$

The matrices $\underline{\mathbf{V}}^{*}_{\text{ut}}(k)$ and $\underline{\mathbf{U}}^{*}_{\text{ap}}(k)$ are also matrices of left and right eigenvectors, respectively, of $\underline{\mathbf{H}}_{\text{cdn}}(k)$. As shown in equations (39) and (40) and based on the above

description, the matrices of left and right eigenvectors for one link are the complex conjugate of the matrices of right and left eigenvectors, respectively, for the other link. The matrices $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$, $\underline{\mathbf{V}}_{\mathrm{ut}}^{*}(k)$, $\underline{\mathbf{V}}_{\mathrm{ut}}^{T}(k)$, and $\underline{\mathbf{V}}_{\mathrm{ut}}^{H}(k)$ are different forms of the matrix $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$, and the matrices $\underline{\mathbf{U}}_{\mathrm{ap}}(k)$, $\underline{\mathbf{U}}_{\mathrm{ap}}^{*}(k)$, $\underline{\mathbf{U}}_{\mathrm{ap}}^{T}(k)$, and $\underline{\mathbf{U}}_{\mathrm{ap}}^{H}(k)$ are also different forms of the matrix $\underline{\mathbf{U}}_{\mathrm{ap}}(k)$. For simplicity, reference to the matrices $\underline{\mathbf{U}}_{\mathrm{ap}}(k)$ and $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$ in the following description may also refer to their various other forms. The matrices $\underline{\mathbf{U}}_{\mathrm{ap}}(k)$ and $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$ are used by the access point and user terminal, respectively, for spatial processing and are denoted as such by their subscripts. The eigenvectors are also often referred to as "steering" vectors.

[00340] The user terminal can estimate the calibrated downlink channel response based on the MIMO pilot sent by the access point. The user terminal can then perform the singular value decomposition of the calibrated downlink channel response estimate $\hat{\underline{\mathbf{H}}}_{\mathrm{cdn}}(k)$, for $k \in K$, to obtain the diagonal matrix $\hat{\underline{\Sigma}}(k)$ and the matrix $\underline{\mathbf{V}}_{\mathrm{ut}}^{*}(k)$ of left eigenvectors of $\hat{\underline{\mathbf{H}}}_{\mathrm{cdn}}(k)$. This singular value decomposition may be given as $\hat{\underline{\mathbf{H}}}_{\mathrm{cdn}}(k) = \hat{\underline{\mathbf{V}}}_{\mathrm{ut}}^{*}(k)\hat{\underline{\Sigma}}(k)\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}^{T}(k)$, where the hat ("$\wedge$") above each matrix indicates that it is an estimate of the actual matrix.

[00341] Similarly, the access point can estimate the calibrated uplink channel response based on a MIMO pilot sent by the user terminal. The access point may then perform singular value decomposition for the calibrated uplink channel response estimate $\hat{\underline{\mathbf{H}}}_{\mathrm{cup}}(k)$, for $k \in K$, to obtain the diagonal matrix $\hat{\underline{\Sigma}}(k)$ and the matrix $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)$ of left eigenvectors of $\hat{\underline{\mathbf{H}}}_{\mathrm{cup}}(k)$. This singular value decomposition may be given as $\hat{\underline{\mathbf{H}}}_{\mathrm{cup}}(k) = \hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)\hat{\underline{\Sigma}}(k)\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}^{H}(k)$.

[00342] An $(N_{ut} \times N_{ut})$ matrix $\underline{\mathbf{F}}_{\mathrm{ut}}(k)$ may be defined as:

$$\underline{\mathbf{F}}_{\mathrm{ut}}(k) = \hat{\underline{\mathbf{K}}}_{\mathrm{ut}}(k)\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k) \quad \text{, for } k \in K . \qquad \text{Eq (41)}$$

While it is active, the user terminal keeps a running estimate of the calibrated downlink channel $\hat{\underline{\mathbf{H}}}_{\mathrm{cdn}}(k)$ and the matrices $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$ of left eigenvectors of $\hat{\underline{\mathbf{H}}}_{\mathrm{cdn}}(k)$, which are used to update the matrix $\underline{\mathbf{F}}_{\mathrm{ut}}(k)$.

[00343] The user terminal uses the matrix $\underline{\mathbf{F}}_{\mathrm{ut}}(k)$ for the spatial processing for the beam-steering and spatial multiplexing modes. For the spatial multiplexing mode, the transmit vector $\underline{\mathbf{x}}_{\mathrm{up}}(k)$ for each subband may be expressed as:

$$\underline{\mathbf{x}}_{\mathrm{up}}(k)=\underline{\mathbf{F}}_{\mathrm{ut}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k) \quad \text{, for } k\in K\text{ ,} \qquad \text{Eq (42)}$$

where $\underline{\mathbf{s}}_{\mathrm{up}}(k)$ is a data vector with up to N_S symbols to be transmitted on the N_S eigenmodes of subband k;

$\underline{\mathbf{F}}_{\mathrm{ut}}(k)$ substitutes for $\underline{\mathbf{V}}(k)$ in equation (15), and the signal scaling by $\underline{\mathbf{G}}(k)$ to achieve channel inversion is omitted in equation (42) for simplicity; and

$\underline{\mathbf{x}}_{\mathrm{up}}(k)$ is the transmit vector for the uplink for subband k.

[00344] At the access point, the received vector $\underline{\mathbf{r}}_{\mathrm{up}}(k)$ for the uplink transmission may be expressed as:

$$\begin{aligned}\underline{\mathbf{r}}_{\mathrm{up}}(k) &= \underline{\mathbf{H}}_{\mathrm{up}}(k)\underline{\mathbf{x}}_{\mathrm{up}}(k)+\underline{\mathbf{n}}_{\mathrm{up}}(k) \qquad \text{, for } k\in K\text{ ,} \qquad \text{Eq (43)}\\ &= \underline{\mathbf{H}}_{\mathrm{up}}(k)\underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k)+\underline{\mathbf{n}}_{\mathrm{up}}(k)\\ &\approx \underline{\hat{\mathbf{H}}}_{\mathrm{cup}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k)+\underline{\mathbf{n}}_{\mathrm{up}}(k)\\ &= \underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)\underline{\hat{\Sigma}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}^{H}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k)+\underline{\mathbf{n}}_{\mathrm{up}}(k)\\ &= \underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)\underline{\hat{\Sigma}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k)+\underline{\mathbf{n}}_{\mathrm{up}}(k)\end{aligned}$$

where $\underline{\mathbf{r}}_{\mathrm{up}}(k)$ is the received vector for the uplink subband k; and

$\underline{\mathbf{n}}_{\mathrm{up}}(k)$ is additive white Gaussian noise (AWGN) for subband k.

Equation (43) uses the following relationships: $\underline{\mathbf{H}}_{\mathrm{up}}(k)\underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k)=\underline{\mathbf{H}}_{\mathrm{cup}}(k)\approx\underline{\hat{\mathbf{H}}}_{\mathrm{cup}}(k)$ and $\underline{\hat{\mathbf{H}}}_{\mathrm{cup}}(k)=\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)\underline{\hat{\Sigma}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}^{H}(k)$. As shown in equation (43), at the access point, the received uplink transmission is transformed by $\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)\underline{\hat{\Sigma}}(k)$, which is the matrix $\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)$ of left eigenvectors of $\underline{\hat{\mathbf{H}}}_{\mathrm{cup}}(k)$ scaled by the diagonal matrix $\underline{\hat{\Sigma}}(k)$ of singular values.

[00345] The user terminal transmits a steered reference on the uplink using the matrix $\underline{\mathbf{F}}_{\mathrm{ut}}(k)$. The steered reference is a pilot transmission on one wideband eigenmode using either beam-steering or beam-forming, and is described in detail below. At the access point, the received uplink steered reference (in the absence of noise) is approximately $\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)\underline{\hat{\Sigma}}(k)$. The access point can thus obtain an estimate of the unitary matrix $\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}(k)$

and diagonal matrix $\hat{\underline{\Sigma}}(k)$ based on the steered reference sent by the user terminal. Various estimation techniques may be used to obtain the estimate of the unitary and diagonal matrices.

[00346] In one embodiment, to obtain an estimate of $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)$, the received vector $\underline{\mathbf{r}}_m(k)$ for the steered reference for subband k of wideband eigenmode m is first multiplied with the complex conjugate of a pilot OFDM symbol, $p^*(k)$, sent for the steered reference. The generation of the steered reference and the pilot OFDM symbol are described in detail below. The result is then integrated over multiple received steered reference symbols for each wideband eigenmode to obtain an estimate of $\hat{\underline{\mathbf{u}}}_m(k)\sigma_m(k)$, which is a scaled left eigenvector of $\hat{\underline{\mathbf{H}}}_{\mathrm{cup}}(k)$ for wideband eigenmode m. Since eigenvectors have unit power, the singular values (or $\sigma_m(k)$) in $\hat{\underline{\Sigma}}(k)$ may be estimated based on the received power of the steered reference, which can be measured for each subband of each wideband eigenmode.

[00347] In another embodiment, an MMSE technique is used to obtain an estimate of $\hat{\underline{\mathbf{u}}}_m(k)$ based on the received vector $\underline{\mathbf{r}}_m(k)$ for the steered reference.

[00348] The steered reference may be sent for one wideband eigenmode in any given symbol period, and may in turn be used to obtain an estimate of one eigenvector for each subband of that wideband eigenmode. Thus, the receiver is able to obtain an estimate of one eigenvector in a unitary matrix for any given symbol period. Since estimates of multiple eigenvectors for the unitary matrix are obtained over different symbol periods, and due to noise and other sources of degradation in the transmission path, the estimated eigenvectors for the unitary matrix are not likely be orthogonal. If the estimated eigenvectors are thereafter used for spatial processing of data transmission on the other link, then any errors in orthogonality in these estimated eigenvectors would result in cross-talk among the eigenmodes, which may degrade performance.

[00349] In an embodiment, the estimated eigenvectors for each unitary matrix are forced to be orthogonal to each other. The orthogonalization of the eigenvectors may be achieved using various techniques such as QR factorization, minimum square error computation, polar decomposition, and so on. QR factorization decomposes a matrix $\underline{\mathbf{M}}^T$ (with non-orthogonal columns) into an orthogonal matrix $\underline{\mathbf{Q}}_{\mathrm{F}}$ and an upper triangle

matrix $\underline{\mathbf{R}}_{\mathrm{F}}$. The matrix $\underline{\mathbf{Q}}_{\mathrm{F}}$ forms an orthogonal basis for the columns of $\underline{\mathbf{M}}^T$. The diagonal elements of $\underline{\mathbf{R}}_{\mathrm{F}}$ give the length of the components of the columns of $\underline{\mathbf{M}}^T$ in the directions of the respective columns of $\underline{\mathbf{Q}}_{\mathrm{F}}$. The matrix $\underline{\mathbf{Q}}_{\mathrm{F}}$ may be used for spatial processing on the downlink. The matrices $\underline{\mathbf{Q}}_{\mathrm{F}}$ and $\underline{\mathbf{R}}_{\mathrm{F}}$ may be used to derive an enhanced matched filter matrix for the uplink. The QR factorization may be performed by various methods including a Gram-Schmidt procedure, a householder transformation, and so on.

[00350] Other techniques to estimate the unitary and diagonal matrices based on the steered reference may also be used, and this is within the scope of the invention.

[00351] The access point can thus estimate both $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)$ and $\hat{\underline{\Sigma}}(k)$ based on the steered reference sent by the user terminal, without having to perform singular value decomposition on $\hat{\underline{\mathbf{H}}}_{\mathrm{cup}}(k)$.

[00352] A normalized matched filter matrix $\underline{\mathbf{M}}_{\mathrm{ap}}(k)$ for the uplink transmission from the user terminal may be expressed as:

$$\underline{\mathbf{M}}_{\mathrm{ap}}(k) = \hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}^{H}(k) \quad \text{, for } k \in K . \qquad \text{Eq (44)}$$

The matched filtering at the access point for the uplink transmission may then be expressed as:

$$\begin{aligned} \hat{\underline{\mathbf{s}}}_{\mathrm{up}}(k) &= \underline{\mathbf{M}}_{\mathrm{ap}}(k)\underline{\mathbf{r}}_{\mathrm{up}}(k) \\ &= \hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}^{H}(k)(\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)\hat{\underline{\Sigma}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k) + \underline{\mathbf{n}}_{\mathrm{up}}(k)) \quad \text{, for } k \in K , \qquad \text{Eq (45)} \\ &= \underline{\mathbf{s}}_{\mathrm{up}}(k) + \tilde{\underline{\mathbf{n}}}_{\mathrm{up}}(k) \end{aligned}$$

where $\hat{\underline{\mathbf{s}}}_{\mathrm{up}}(k)$ is an estimate of the vector of modulation symbols $\underline{\mathbf{s}}_{\mathrm{up}}(k)$ transmitted by the user terminal for the spatial multiplexing mode. For the beam-steering mode, only one row of the matrix $\underline{\mathbf{M}}_{\mathrm{ap}}(k)$ is used to provide one symbol estimate $\hat{s}(k)$ for the eigenmode used for data transmission.

2. Downlink Spatial Processing

[00353] For the downlink, the access point uses an $(N_{ap} \times N_{ap})$ matrix $\underline{\mathbf{F}}_{\mathrm{ap}}(k)$ for spatial processing. This matrix may be expressed as:

$$\underline{\mathbf{F}}_{\mathrm{ap}}(k) = \hat{\underline{\mathbf{K}}}_{\mathrm{ap}}(k)\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}^{*}(k) \quad \text{, for } k \in K . \qquad \text{Eq (46)}$$

The correction matrix $\hat{\underline{\mathbf{K}}}_{ap}(k)$ is derived by the user terminal and sent back to the access point during calibration. The matrix $\hat{\underline{\mathbf{U}}}_{ap}(k)$ may be obtained based on the steered reference sent on the uplink by the user terminal.

[00354] For the spatial multiplexing mode, the transmit vector $\underline{\mathbf{x}}_{dn}(k)$ for the downlink for each data subband may be expressed as:

$$\underline{\mathbf{x}}_{dn}(k) = \underline{\mathbf{F}}_{ap}(k)\underline{\mathbf{s}}_{dn}(k) \quad \text{, for } k \in K \text{,} \qquad \text{Eq (47)}$$

where $\underline{\mathbf{x}}_{dn}(k)$ is the transmit vector, $\underline{\mathbf{s}}_{dn}(k)$ is the data vector for the downlink, and the signal scaling by $\underline{\mathbf{G}}(k)$ to achieve channel inversion is again omitted for simplicity.

[00355] At the user terminal, the received vector $\underline{\mathbf{r}}_{dn}(k)$ for the downlink transmission may be expressed as:

$$\begin{aligned}
\underline{\mathbf{r}}_{dn}(k) &= \underline{\mathbf{H}}_{dn}(k)\underline{\mathbf{x}}_{dn}(k) + \underline{\mathbf{n}}_{dn}(k) \\
&= \underline{\mathbf{H}}_{dn}(k)\hat{\underline{\mathbf{K}}}_{ap}(k)\hat{\underline{\mathbf{U}}}^{*}_{ap}(k)\underline{\mathbf{s}}_{dn}(k) + \underline{\mathbf{n}}_{dn}(k) \\
&\approx \hat{\underline{\mathbf{H}}}_{cdn}(k)\hat{\underline{\mathbf{U}}}^{*}_{ap}(k)\underline{\mathbf{s}}_{dn}(k) + \underline{\mathbf{n}}_{dn}(k) \\
&= \hat{\underline{\mathbf{V}}}^{*}_{ut}(k)\hat{\underline{\Sigma}}(k)\hat{\underline{\mathbf{U}}}^{T}_{ap}(k)\hat{\underline{\mathbf{U}}}^{*}_{ap}(k)\underline{\mathbf{s}}_{dn}(k) + \underline{\mathbf{n}}_{dn}(k)) \\
&= \hat{\underline{\mathbf{V}}}^{*}_{ut}(k)\hat{\underline{\Sigma}}(k)\underline{\mathbf{s}}_{dn}(k) + \underline{\mathbf{n}}_{dn}(k) \qquad \text{, for } k \in K \text{.} \quad \text{Eq (48)}
\end{aligned}$$

As shown in equation (48), at the user terminal, the received downlink transmission is transformed by $\hat{\underline{\mathbf{V}}}^{*}_{ut}(k)\hat{\underline{\Sigma}}(k)$, which is the matrix $\hat{\underline{\mathbf{V}}}_{ut}(k)$ of left eigenvectors of $\hat{\underline{\mathbf{H}}}_{cdn}(k)$ scaled by the diagonal matrix $\hat{\underline{\Sigma}}(k)$ of singular values.

[00356] A normalized matched filter matrix $\underline{\mathbf{M}}_{ut}(k)$ for the downlink transmission from the access point may be expressed as:

$$\underline{\mathbf{M}}_{ut}(k) = \hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{V}}}^{T}_{ut}(k) \quad \text{, for } k \in K \text{.} \qquad \text{Eq (49)}$$

The diagonal matrix $\hat{\underline{\Sigma}}(k)$ and matrix $\hat{\underline{\mathbf{V}}}_{ut}(k)$ of left eigenvectors can be derived by the user terminal by performing singular value decomposition on the calibrated downlink channel response estimate $\hat{\underline{\mathbf{H}}}_{cdn}(k)$, as described above.

[00357] The matched filtering at the user terminal for the downlink transmission may then be expressed as:

$$\begin{aligned}\hat{\underline{\mathbf{s}}}_{\text{dn}}(k) &= \underline{\mathbf{M}}_{\text{ut}}(k)\underline{\mathbf{r}}_{\text{dn}}(k) \\ &= \hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{V}}}_{\text{ut}}^{T}(k)(\hat{\underline{\mathbf{V}}}_{\text{ut}}^{*}(k)\hat{\underline{\Sigma}}(k)\underline{\mathbf{s}}_{\text{dn}}(k)+\underline{\mathbf{n}}_{\text{dn}}(k)) \quad \text{, for } k \in K. \\ &= \underline{\mathbf{s}}_{\text{dn}}(k)+\tilde{\underline{\mathbf{n}}}_{\text{dn}}(k)\end{aligned} \qquad \text{Eq (50)}$$

3. Access Point and User Terminal Spatial Processing

[00358] Because of the reciprocal channel for the TDD system and the calibration, the spatial processing at both the access point and the user terminal may be simplified. Table 32 summarizes the spatial processing at the access point and user terminal for data transmission and reception.

Table 32

	Uplink	**Downlink**
User Terminal	Transmit : $\underline{\mathbf{x}}_{\text{up}}(k)=\hat{\underline{\mathbf{K}}}_{\text{ut}}(k)\hat{\underline{\mathbf{V}}}_{\text{ut}}(k)\underline{\mathbf{s}}_{\text{up}}(k)$	Receive : $\hat{\underline{\mathbf{s}}}_{\text{dn}}(k)=\hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{V}}}_{\text{ut}}^{T}(k)\underline{\mathbf{r}}_{\text{dn}}(k)$
Access Point	Receive : $\hat{\underline{\mathbf{s}}}_{\text{up}}(k)=\hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{U}}}_{\text{ap}}^{H}(k)\underline{\mathbf{r}}_{\text{up}}(k)$	Transmit : $\underline{\mathbf{x}}_{\text{dn}}(k)=\hat{\underline{\mathbf{K}}}_{\text{ap}}(k)\hat{\underline{\mathbf{U}}}_{\text{ap}}^{*}(k)\underline{\mathbf{s}}_{\text{dn}}(k)$

The spatial processing for data reception is also commonly referred to as matched filtering.

[00359] Because of the reciprocal channel, $\hat{\underline{\mathbf{V}}}_{\text{ut}}(k)$ is both the matrix of right eigenvectors of $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$ (to transmit) and left eigenvectors of $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$ (to receive) for the user terminal. Similarly, $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ is both the matrix of right eigenvectors of $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$ (to transmit) and left eigenvectors of $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$ (to receive) for the access point. The singular value decomposition only needs to be performed by the user terminal for the calibrated downlink channel response estimate $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$ to obtain $\hat{\underline{\mathbf{V}}}_{\text{ut}}(k)$ and $\hat{\underline{\Sigma}}(k)$. The access point can derive $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ and $\hat{\underline{\Sigma}}(k)$ based on the steered reference sent by the user terminal and does not need to perform the singular value decomposition on the uplink channel response estimate $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$. The access point and user terminal may have different versions of the matrix $\hat{\underline{\Sigma}}(k)$ due to the different means used by the access point and user terminal to derive $\hat{\underline{\Sigma}}(k)$. Moreover, the matrix $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ derived by the

access point based on the steered reference is typically different from the matrix $\hat{\underline{\mathbf{U}}}_{ap}(k)$ derived by the user terminal using singular value decomposition. For simplicity, these differences are not shown in the above derivation.

4. Beam-Steering

[00360] For certain channel conditions, it is better to transmit data on only one wideband eigenmode - typically the best or principal wideband eigenmode. This may be the case if the received SNRs for all other wideband eigenmodes are sufficiently poor so that improved performance is achieved by using all of the available transmit power on the principal wideband eigenmode.

[00361] Data transmission on one wideband eigenmode may be achieved using either beam-forming or beam-steering. For beam-forming, the modulation symbols are spatially processed with the eigenvectors $\hat{\underline{\mathbf{v}}}_{ut,1}(k)$ or $\hat{\underline{\mathbf{u}}}_{ap,1}(k)$, for $k \in K$, for the principal wideband eigenmode (i.e., the first column of $\hat{\underline{\mathbf{V}}}_{ut}(k)$ or $\hat{\underline{\mathbf{U}}}_{ap}(k)$, after the ordering). For beam-steering, the modulation symbols are spatially processed with a set of "normalized" (or "saturated") eigenvectors $\tilde{\underline{\mathbf{v}}}_{ut}(k)$ or $\tilde{\underline{\mathbf{u}}}_{ap}(k)$, for $k \in K$, for the principal wideband eigenmode. For clarity, beam-steering is described below for the uplink.

[00362] For the uplink, the elements of each eigenvector $\hat{\underline{\mathbf{v}}}_{ut,1}(k)$, for $k \in K$, for the principal wideband eigenmode may have different magnitudes. Thus, the preconditioned symbols for each subband, which are obtained by multiplying the modulation symbol for subband k with the elements of the eigenvector $\hat{\underline{\mathbf{v}}}_{ut,1}(k)$ for subband k, may then have different magnitudes. Consequently, the per-antenna transmit vectors, each of which includes the preconditioned symbols for all data subbands for a given transmit antenna, may have different magnitudes. If the transmit power for each transmit antenna is limited (e.g., because of limitations of power amplifiers), then beam-forming may not fully use the total power available for each antenna.

[00363] Beam-steering uses only the phase information from the eigenvectors $\hat{\underline{\mathbf{v}}}_{ut,1}(k)$, for $k \in K$, for the principal wideband eigenmode and normalizes each eigenvector such that all elements in the eigenvector have equal magnitudes. The normalized eigenvector $\tilde{\underline{\mathbf{v}}}_{ut}(k)$ for subband k may be expressed as:

$$\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k) = [Ae^{j\theta_1(k)} \;\; Ae^{j\theta_2(k)} \;\; \ldots \;\; Ae^{j\theta_{N_{ut}}(k)}]^T \quad , \qquad \text{Eq (51)}$$

where A is a constant (e.g., $A = 1$); and

$\theta_i(k)$ is the phase for subband k of transmit antenna i, which is given as:

$$\theta_i(k) = \angle \hat{v}_{\text{ut},1,i}(k) = \tan^{-1}\left(\frac{\text{Im}\{\hat{v}_{\text{ut},1,i}(k)\}}{\text{Re}\{\hat{v}_{\text{ut},1,i}(k)\}}\right) . \qquad \text{Eq (52)}$$

As shown in equation (52), the phase of each element in the vector $\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k)$ is obtained from the corresponding element of the eigenvector $\underline{\hat{\mathbf{v}}}_{\text{ut},1}(k)$ (i.e., $\theta_i(k)$ is obtained from $\hat{v}_{\text{ut},1,i}(k)$, where $\underline{\hat{\mathbf{v}}}_{\text{ut},1}(k) = [\hat{v}_{\text{ut},1,1}(k) \;\; \hat{v}_{\text{ut},1,2}(k) \;\; \ldots \;\; \hat{v}_{\text{ut},1,N_{ut}}(k)]^T$).

5. Uplink Beam-Steering

[00364] The spatial processing by the user terminal for beam-steering on the uplink may be expressed as:

$$\underline{\tilde{\mathbf{x}}}_{\text{up}}(k) = \underline{\hat{\mathbf{K}}}_{\text{ut}}\,\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k)s_{\text{up}}(k) \quad , \text{for } k \in K , \qquad \text{Eq (53)}$$

where $s_{\text{up}}(k)$ is the modulation symbol to be transmitted on subband k; and

$\underline{\tilde{\mathbf{x}}}_{\text{up}}(k)$ is the transmit vector for subband k for beam-steering.

As shown in equation (53), the N_{ut} elements of the normalized steering vector $\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k)$ for each subband have equal magnitude but possibly different phases. The beam-steering thus generates one transmit vector $\underline{\tilde{\mathbf{x}}}_{\text{up}}(k)$ for each subband, with the N_{ut} elements of $\underline{\tilde{\mathbf{x}}}_{\text{up}}(k)$ having the same magnitude but possibly different phases.

[00365] The received uplink transmission at the access point for beam-steering may be expressed as:

$$\begin{aligned} \underline{\tilde{\mathbf{r}}}_{\text{up}}(k) &= \underline{\mathbf{H}}_{\text{up}}(k)\underline{\tilde{\mathbf{x}}}_{\text{up}}(k) + \underline{\mathbf{n}}_{\text{up}}(k) \qquad , \text{for } k \in K , \qquad \text{Eq (54)} \\ &= \underline{\mathbf{H}}_{\text{up}}(k)\underline{\hat{\mathbf{K}}}_{\text{ut}}(k)\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k)s_{\text{up}}(k) + \underline{\mathbf{n}}_{\text{up}}(k) \\ &= \underline{\mathbf{H}}_{\text{cup}}(k)\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k)s_{\text{up}}(k) + \underline{\mathbf{n}}_{\text{up}}(k) \end{aligned}$$

where $\underline{\tilde{\mathbf{r}}}_{\text{up}}(k)$ is the received vector for the uplink for subband k for beam-steering.

[00366] A matched filter row vector $\underline{\tilde{\mathbf{m}}}_{\text{ap}}(k)$ for the uplink transmission using beam-steering may be expressed as:

$$\underline{\tilde{\mathbf{m}}}_{\text{ap}}(k) = (\underline{\mathbf{H}}_{\text{cup}}(k)\underline{\tilde{\mathbf{v}}}_{\text{ut}}(k))^H , \text{ for } k \in K . \qquad \text{Eq (55)}$$

The matched filter vector $\tilde{\underline{\mathbf{m}}}_{\text{ap}}(k)$ may be obtained as described below. The spatial processing (or matched filtering) at the access point for the received uplink transmission with beam-steering may be expressed as:

$$\begin{aligned}\hat{s}_{\text{up}}(k) &= \tilde{\lambda}_{\text{up}}^{-1}(k)\tilde{\underline{\mathbf{m}}}_{\text{ap}}(k)\tilde{\underline{\mathbf{r}}}_{\text{up}}(k)\\ &= \tilde{\lambda}_{\text{up}}^{-1}(k)(\underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k))^H(\underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)s_{\text{up}}(k)+\underline{\mathbf{n}}_{\text{up}}(k)) \text{ , for } k \in K\text{ ,} \qquad \text{Eq (56)}\\ &= s_{\text{up}}(k)+\tilde{n}_{\text{up}}(k)\end{aligned}$$

where $\tilde{\lambda}_{\text{up}}(k)=(\underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k))^H(\underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k))$ (i.e., $\tilde{\lambda}_{\text{up}}(k)$ is the inner product of $\tilde{\underline{\mathbf{m}}}_{\text{ap}}(k)$ and its conjugate transpose),

$\hat{s}_{\text{up}}(k)$ is an estimate of the modulation symbol $s_{\text{up}}(k)$ transmitted by the user terminal on the uplink, and

$\tilde{n}_{\text{up}}(k)$ is the post-processed noise.

6. Downlink Beam-Steering

[00367] The spatial processing by the access point for beam-steering on the downlink may be expressed as:

$$\tilde{\underline{\mathbf{x}}}_{\text{dn}}(k)=\hat{\underline{\mathbf{K}}}_{\text{ap}}\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)s_{\text{dn}}(k) \text{ , for } k \in K\text{ ,} \qquad \text{Eq (57)}$$

where $\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$ is the normalized eigenvector for subband k, which is generated based on the eigenvector $\hat{\underline{\mathbf{u}}}^*_{\text{ap},1}(k)$, for the principal wideband eigenmode, similar to that described above for the uplink.

[00368] A matched filter row vector $\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)$ for the downlink transmission using beam-steering may be expressed as:

$$\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)=(\underline{\mathbf{H}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k))^H \text{ , for } k \in K\text{ .} \qquad \text{Eq (58)}$$

The spatial processing (or matched filtering) at the user terminal for the received downlink transmission may be expressed as:

$$\begin{aligned}\hat{s}_{\text{dn}}(k) &= \tilde{\lambda}_{\text{dn}}^{-1}(k)\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)\tilde{\underline{\mathbf{r}}}_{\text{dn}}(k)\\ &= \tilde{\lambda}_{\text{dn}}^{-1}(k)(\underline{\mathbf{H}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k))^H(\underline{\mathbf{H}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)s_{\text{up}}(k)+\underline{\mathbf{n}}_{\text{dn}}(k))\text{, for } k \in K\text{ ,} \qquad \text{Eq (59)}\\ &= s_{\text{dn}}(k)+\tilde{n}_{\text{dn}}(k)\end{aligned}$$

where $\tilde{\lambda}_{\text{dn}}(k)=(\underline{\mathbf{H}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k))^H(\underline{\mathbf{H}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k))$ (i.e., $\tilde{\lambda}_{\text{dn}}(k)$ is the inner product of $\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)$ and its conjugate transpose).

7. Spatial Processing with Channel Inversion

[00369] For the uplink, the transmit vector $\underline{\mathbf{x}}_{\mathrm{up}}(k)$ for the spatial multiplexing mode may be derived by the user terminal as:

$$\underline{\mathbf{x}}_{\mathrm{up}}(k) = \underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}(k)\underline{\mathbf{G}}(k)\underline{\mathbf{s}}_{\mathrm{up}}(k) \quad \text{, for } k \in K \text{ ,} \qquad \text{Eq (60)}$$

where $\underline{\mathbf{G}}(k)$ is a diagonal matrix of gains for the channel inversion described above. Equation (60) is similar to equation (15), except that $\underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}(k)$ is used in place of $\underline{\mathbf{V}}(k)$. The elements of $\underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k)\underline{\hat{\mathbf{V}}}_{\mathrm{ut}}(k)$ are provided to multipliers 952 within beam-formers 950 in FIG. 9B.

[00370] For the uplink, the transmit vector $\underline{\tilde{\mathbf{x}}}_{\mathrm{up}}(k)$ for the beam-steering mode may be derived by the user terminal as:

$$\underline{\tilde{\mathbf{x}}}_{\mathrm{up}}(k) = \underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k)\underline{\tilde{\mathbf{v}}}_{\mathrm{ut}}(k)\tilde{g}(k)s_{\mathrm{up}}(k) \quad \text{, for } k \in K \text{ ,} \qquad \text{Eq (61)}$$

where $\underline{\tilde{\mathbf{v}}}_{\mathrm{ut}}(k)$ is a vector with four elements having equal magnitude but phases obtained based on eigenvector $\underline{\hat{\mathbf{v}}}_{\mathrm{ut},1}(k)$ for the principal eigenmode. The vector $\underline{\tilde{\mathbf{v}}}_{\mathrm{ut}}(k)$ may be derived similar to that shown above in equations (16) and (17). The gain $\tilde{g}(k)$ achieves channel inversion and may be derived similar to that shown above in equations (18) through (20), except that $\tilde{\lambda}_1(k) = \underline{\tilde{\mathbf{v}}}_{\mathrm{ut}}^H(k)\underline{\hat{\mathbf{H}}}_{\mathrm{cup}}^H(k)\underline{\hat{\mathbf{H}}}_{\mathrm{cup}}(k)\underline{\tilde{\mathbf{v}}}_{\mathrm{ut}}(k)$ is used for equation (20). The elements of $\underline{\tilde{\mathbf{v}}}_{\mathrm{ut}}(k)$ are provided to multipliers 1052 within beam-steering unit 1050 in FIG. 10B.

[00371] For the downlink, the transmit vector $\underline{\mathbf{x}}_{\mathrm{dn}}(k)$ for the spatial multiplexing mode may be derived by the access point as:

$$\underline{\mathbf{x}}_{\mathrm{dn}}(k) = \underline{\hat{\mathbf{K}}}_{\mathrm{ap}}(k)\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}^{*}(k)\underline{\mathbf{G}}(k)\underline{\mathbf{s}}_{\mathrm{dn}}(k) \quad \text{, for } k \in K \text{.} \qquad \text{Eq (62)}$$

Equation (62) is similar to equation (15), except that $\underline{\hat{\mathbf{K}}}_{\mathrm{ap}}(k)\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}^{*}(k)$ is used in place of $\underline{\mathbf{V}}(k)$. The elements of $\underline{\hat{\mathbf{K}}}_{\mathrm{ap}}(k)\underline{\hat{\mathbf{U}}}_{\mathrm{ap}}^{*}(k)$ may be provided to multipliers 952 within beam-formers 950 in FIG. 9B.

[00372] For the downlink, the transmit vector $\underline{\tilde{\mathbf{x}}}_{\mathrm{dn}}(k)$ for the beam-steering mode may be derived by the access point as:

$$\underline{\tilde{\mathbf{x}}}_{\mathrm{dn}}(k) = \underline{\hat{\mathbf{K}}}_{\mathrm{ap}}(k)\underline{\tilde{\mathbf{u}}}_{\mathrm{ap}}(k)\tilde{g}(k)s_{\mathrm{dn}}(k) \quad \text{, for } k \in K \text{ ,} \qquad \text{Eq (63)}$$

where $\underline{\tilde{\mathbf{u}}}_{\text{ap}}(k)$ is a vector with four elements having equal magnitude but phases obtained based on eigenvector $\underline{\hat{\mathbf{u}}}_{\text{ap},1}(k)$ for the principal eigenmode. The gain $\tilde{g}(k)$ achieves channel inversion and may be derived in a similar manner to that shown above in equations (18) through (20), except that $\tilde{\lambda}_1(k) = \underline{\tilde{\mathbf{u}}}_{\text{ap}}^H(k)\underline{\hat{\mathbf{H}}}_{\text{cdn}}^H(k)\underline{\hat{\mathbf{H}}}_{\text{cdn}}(k)\underline{\tilde{\mathbf{u}}}_{\text{ap}}(k)$ is used for equation (20). The elements of $\underline{\tilde{\mathbf{u}}}_{\text{ap}}(k)$ are provided to multipliers 1052 within beam-steering unit 1050 in FIG. 10B.

VIII. Pilot Structure

[00373] A pilot structure is provided for the MIMO WLAN system to allow the access points and user terminals to perform timing and frequency acquisition, channel estimation, and other functions needed for proper system operation. Table 33 lists four types of pilot and their short description for an exemplary pilot structure. Fewer, different, and/or additional pilot types may also be used for the pilot structure.

Table 33 - Pilot Types

Pilot Type	Description
Beacon Pilot	A pilot transmitted from all transmit antennas and used for timing and frequency acquisition.
MIMO Pilot	A pilot transmitted from all transmit antennas with different orthogonal codes and used for channel estimation.
Steered Reference or Steered Pilot	A pilot transmitted on specific eigenmodes of a MIMO channel for a specific user terminal and used for channel estimation and possibly rate control.
Carrier Pilot	A pilot used for phase tracking of a carrier signal.

Steered reference and steered pilot are synonymous terms.

[00374] In an embodiment, the pilot structure includes (1) for the downlink - a beacon pilot, a MIMO pilot, a steered reference, and a carrier pilot transmitted by the access point, and (2) for the uplink - a MIMO pilot, a steered reference, and a carrier pilot transmitted by the user terminals.

[00375] The downlink beacon pilot and MIMO pilot are sent on the BCH (as shown in FIG. 5A) in each TDD frame. The beacon pilot may be used by the user terminals for timing and frequency acquisition and Doppler estimation. The MIMO pilot may be

used by the user terminals to (1) obtain an estimate of the downlink MIMO channel, (2) derive the steering vectors for uplink transmission (if the beam-steering or spatial multiplexing mode is supported), and (3) derive a matched filter for downlink transmission. The downlink steered reference may also be used by a specified user terminal for channel estimation.

[00376] An uplink steered reference is transmitted by each active user terminal that supports the beam-steering or spatial multiplexing mode and may be used by the access point to (1) derive the steering vectors for the downlink transmission and (2) derive a matched filter for the uplink transmission. In general, the steered reference is only sent for/by user terminals that support the beam-steering and/or spatial multiplexing modes. The reference sent works regardless of whether or not it is steered properly (e.g., due to a poor channel estimate). That is, the reference becomes orthogonal on a per transmit antenna basis since the steering matrix is diagonal.

[00377] If a user terminal is calibrated, then it can transmit a steered reference on the principal eigenmode on the RACH using the vector $\hat{\underline{\mathbf{K}}}_{\text{ut}}(k)\hat{\underline{\mathbf{v}}}_{\text{ut},0}(k)$, for $k \in K$, where $\hat{\underline{\mathbf{v}}}_{\text{ut},0}(k)$ is the column of $\hat{\underline{\mathbf{V}}}_{\text{ut}}(k)$ for the principal eigenmode. If the user terminal is not calibrated, then it can transmit a pilot on the RACH using a vector $\underline{\mathbf{v}}_{\text{ut,p}}(k) = [e^{j\theta_1(k)} \; e^{j\theta_2(k)} \; \ldots \; e^{j\theta_{N_{ut}}(k)}]^T$, for $k \in K$. The vector $\underline{\mathbf{v}}_{\text{ut,p}}(k)$ for each subband includes N_{ut} random steering coefficients whose phases $\theta_i(k)$, for $i \in \{1, 2, \ldots N_{ut}\}$, may be selected in accordance with a pseudo-random procedure. Since only the relative phases among the N_{ut} steering coefficients matter, the phase of the first steering coefficient may be set to zero (i.e., $\theta_1(k) = 0$). The phases of the other $N_{ut} - 1$ steering coefficients may change for each access attempt, so that all 360° are covered by each steering coefficient in intervals of $360° / N_{\theta_i}$, where N_{θ_i} is a function of N_{ut}. The perturbation of the phases of the N_{ut} elements of the steering vector $\underline{\mathbf{v}}_{\text{ut,p}}(k)$ on every RACH attempt, when using the RACH in the beam-steering mode prior to calibration, is so that the user terminal does not use a bad steering vector for all access attempts. A MIMO pilot may be sent for/by user terminals that do not support beam-steering and/or spatial multiplexing modes.

[00378] The access point does not have knowledge of the channel for any user terminal until the user terminal communicates directly with the access point. When a user

terminal desires to transmit data, it first estimates the channel based on the MIMO pilot transmitted by the access point. The user terminal then sends steered reference in the preamble of the RACH when it attempts to access the system. The access point uses the reference on the RACH preamble for signal detection and channel estimation.

[00379] Once the user terminal has been granted access to the system and assigned FCH/RCH resources by the access point, the user terminal sends a reference (e.g., a MIMO pilot) at the beginning of each RCH PDU it transmits. If the user terminal is using the diversity mode, then the reference is sent on the RCH without steering. If the user terminal is using the beam-steering or spatial multiplexing mode, then a steered reference is sent on the RCH to allow the access point to determine the eigenvector for the principal eigenmode (for the beam-steering mode) or the set of four eigenvectors (for the spatial multiplexing mode) for each of the 48 data subbands. The steered reference allows the access point to improve its estimate of the channel and to track the channel.

1. <u>Beacon Pilot - Downlink</u>

[00380] The downlink beacon pilot is included in the first portion of the BCH (as shown in FIG. 5A) and transmitted in each TDD frame. The beacon pilot includes a specific OFDM symbol (denoted as "B") that is transmitted from each of the four antennas at the access point. The same B OFDM symbol is transmitted twice in the 2-symbol duration for the beacon pilot.

[00381] In a specific embodiment, the B OFDM symbol comprises a set of 12 BPSK modulation symbols, $b(k)$, for 12 specific subbands, which is shown in Table 34.

Table 34 - Pilot Symbols

Sub-band Index	Beacon Pilot $b(k)$	MIMO Pilot $p(k)$	Sub-band Index	Beacon Pilot $b(k)$	MIMO Pilot $p(k)$	Sub-band Index	Beacon Pilot $b(k)$	MIMO Pilot $p(k)$	Sub-band Index	Beacon Pilot $b(k)$	MIMO Pilot $p(k)$
K	0	0	-13	0	$1-j$	1	0	$1-j$	15	0	$1+j$
-26	0	$-1-j$	-12	$-1-j$	$1-j$	2	0	$-1-j$	16	$1+j$	$-1+j$
-25	0	$-1+j$	-11	0	$-1-j$	3	0	$-1-j$	17	0	$-1+j$
-24	$1+j$	$-1+j$	-10	0	$-1-j$	4	$-1-j$	$-1-j$	18	0	$1-j$
-23	0	$-1+j$	-9	0	$1-j$	5	0	$-1+j$	19	0	$1+j$
-22	0	$1-j$	-8	$-1-j$	$-1-j$	6	0	$1+j$	20	$1+j$	$-1+j$

-21	0	1-j	-7	0	1+j	7	0	-1-j	21	0	1+j
-20	-1-j	1+j	-6	0	-1+j	8	-1-j	-1+j	22	0	-1+j
-19	0	-1-j	-5	0	-1-j	9	0	-1-j	23	0	1+j
-18	0	-1+j	-4	1+j	-1+j	10	0	-1-j	24	1+j	-1+j
-17	0	1+j	-3	0	-1+j	11	0	1+j	25	0	1-j
-16	1+j	-1+j	-2	0	1-j	12	1+j	1-j	26	0	-1-j
-15	0	1-j	-1	0	-1+j	13	0	-1+j	\	0	0
-14	0	1+j	0	0	0	14	0	-1-j			

[00382] For the beacon pilot embodiment shown in Table 34, the B OFDM symbol comprises (1) BPSK modulation symbol $(1+j)$ for subbands -24, -16, -4, 12, 16, 20, and 24, (2) BPSK modulation symbol $-(1+j)$ for subbands -20, -12, -8, 4, and 8, and (3) signal values of zero for the remaining 52 subbands. The B OFDM symbol is specifically designed to facilitate timing and frequency acquisition by the user terminals. However, other OFDM symbols may also be used for the beacon pilot, and this is within the scope of the invention.

2. MIMO Pilot - Downlink

[00383] The downlink MIMO pilot is included in the second portion of the BCH (as shown in FIG. 5A) and also transmitted in each TDD frame. The MIMO pilot includes a specific OFDM symbol (denoted as "P") that is transmitted from each of the four antennas at the access point. The same P OFDM symbol is transmitted eight times in the 8-symbol duration for the MIMO pilot. However, the eight P OFDM symbols for each antenna are "covered" with a different 4-chip Walsh sequence assigned to that antenna. Covering is a process whereby a given pilot or data symbol (or a set of L pilot/data symbols with the same value) to be transmitted is multiplied by all L chips of an L-chip orthogonal sequence to obtain L covered symbols, which are then transmitted. Decovering is a complementary process whereby received symbols are multiplied by the L chips of the same L-chip orthogonal sequence to obtain L decovered symbols, which are then accumulated to obtain an estimate of the transmitted pilot/data symbol. The covering achieves orthogonality among the N_T pilot transmissions from the N_T transmit antennas and allows the user terminals to distinguish the individual transmit

antennas. Covering may be achieved with Walsh sequences or other orthogonal sequences.

[00384] In a specific embodiment, the P OFDM symbol comprises a set of 52 QPSK modulation symbols, $p(k)$, for the 48 data subbands and 4 pilot subbands, which is shown in Table 34. Signal values of zero are transmitted on the remaining 12 subbands. The P OFDM symbol comprises a unique "word" of 52 QPSK modulation symbols that is designed to facilitate channel estimation by the user terminals. This unique word is also selected to minimize the peak-to-average variation in the transmitted MIMO pilot. This may then reduce the amount of distortion and non-linearity generated by the receiver circuitry at the user terminals, which can result in improved accuracy for the channel estimation. However, other OFDM symbols may also be used for the MIMO pilot, and this is within the scope of the invention.

[00385] In an embodiment, the four antennas at the access point are assigned 4-chip Walsh sequences of $W_1 = 1111$, $W_2 = 1010$, $W_3 = 1100$, and $W_4 = 1001$ for the MIMO pilot. For a given Walsh sequence, a value of "1" indicates that a P OFDM symbol is transmitted and a value of "0" indicates that a -P OFDM symbol is transmitted (i.e., each of the 52 modulation symbols in P is inverted).

[00386] Table 35 lists the OFDM symbols to be transmitted from each of the four antennas at the access point for the beacon pilot and MIMO pilot. The B and P OFDM symbols are as described above.

Table 35 - Beacon and MIMO Pilots

Pilot	**OFDM Symbol**	**Antenna 1**	**Antenna 2**	**Antenna 3**	**Antenna 4**
Beacon Pilot	1	B	B	B	B
	2	B	B	B	B
MIMO Pilot	3	+P	+P	+P	+P
	4	+P	-P	+P	-P
	5	+P	+P	-P	-P
	6	+P	-P	-P	+P
	7	+P	+P	+P	+P
	8	+P	-P	+P	-P
	9	+P	+P	-P	-P

	10	+P	-P	-P	+P

[00387] The MIMO pilot may be used by the user terminal to estimate the channel response of the downlink. In particular, to recover the pilot sent from access point antenna i and received by user terminal antenna j, the pilot received by terminal antenna j is first multiplied with the Walsh sequence assigned to access point antenna i. The eight decovered OFDM symbols for all eight symbol periods for the MIMO pilot are then accumulated, where the accumulation is performed individually for each of the 52 subbands used to carry the MIMO pilot. The results of the accumulation is $\hat{h}_{\text{cdn}\,i,j}(k)$, for $k = \pm\{1, ..., 26\}$, which is an estimate of the calibrated downlink channel response from access point antenna i to user terminal antenna j for the 52 data and pilot subbands.

[00388] The same pilot processing may be performed by the access point to recover the pilot from each access point antenna at each user terminal antenna. The pilot transmitted from each access point antenna may be recovered by decovering with the Walsh sequence assigned to that antenna. The pilot processing provides $N_{ap} \cdot N_{ut}$ values for each of the 52 subbands, where N_{ap} denotes the number of antennas at the access point and N_{ut} denotes the number of antennas at the user terminal. The $N_{ap} \cdot N_{ut}$ values for each subband are the elements of the calibrated downlink channel response estimate $\underline{\hat{\mathbf{H}}}_{\text{cdn}}(k)$ for that subband.

[00389] The MIMO pilot may also be transmitted on the uplink by the user terminal for calibration and in the diversity mode. The same processing described above for the user terminal to recover the MIMO pilot sent by the access point may also be performed by the access point to recover the MIMO pilot sent by the user terminal.

3. Steered Reference

[00390] A steered reference may be transmitted in the preamble portion of an RACH PDU (as shown in FIG. 5C) or an RCH PDU (as shown in FIGS. 5E and 5G) by each active user terminal. A steered reference may also be transmitted in the preamble portion of an FCH PDU (as shown in FIGS. 5E and 5F) by the access point to an active user terminal.

A. Steered Reference for Spatial Multiplexing

[00391] The steered reference comprises a specific OFDM symbol (e.g., the same P OFDM symbol used for the MIMO pilot) that is transmitted from all of the transmit antennas at the user terminal (for the uplink) or the access point (for the downlink). However, the P OFDM symbol for each symbol period is spatially processed (i.e., beam-formed) with a steering vector for one eigenmode.

[00392] The first symbol of steered reference transmitted by the user terminal in the preamble of the RACH may be expressed as:

$$\underline{\mathbf{x}}(k) = \hat{\underline{\mathbf{K}}}_{\mathrm{ut}}(k) \cdot \hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k) \cdot p(k) \quad \text{, for } k \in K', \qquad \text{Eq (64)}$$

where $\underline{\mathbf{x}}(k)$ is the transmit vector for subband k;

$\hat{\underline{\mathbf{K}}}_{\mathrm{ut}}(k)$ is the correction matrix for subband k for the user terminal;

$\hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k)$ is the steering vector for subband k of the principal wideband eigenmode;

$p(k)$ is the pilot symbol for subband k; and

$K' = \{-32, \ldots\ldots, 31\}$ is the set of indices for all 64 subbands.

The vector $\underline{\mathbf{x}}(k)$ includes four transmit symbols for each value of k, which are to be transmitted from the four antennas at the user terminal. The steering vector $\hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k)$ is the first column of the matrix $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$ of right eigenvectors of the calibrated uplink channel response estimate $\hat{\underline{\mathbf{H}}}_{\mathrm{cup}}(k)$, where $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k) = [\hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k)\ \hat{\underline{\mathbf{v}}}_{\mathrm{ut},2}(k)\ \hat{\underline{\mathbf{v}}}_{\mathrm{ut},3}(k)\ \hat{\underline{\mathbf{v}}}_{\mathrm{ut},4}(k)]$ and $\hat{\underline{\mathbf{v}}}_{\mathrm{ut},i}(k)$ is the i-th column of $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$. The above assumes that the singular values in $\hat{\underline{\Sigma}}(k)$ and the columns of $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$ are ordered as described above.

[00393] The second symbol of steered reference transmitted by the user terminal in the preamble of the RACH includes the data rate indicator (DRI) for the RACH PDU. The DRI indicates the rate used for the RACH message sent in the RACH PDU. The DRI is embedded in the second steered reference symbol by mapping the DRI to a specific QPSK symbol s_{dri}, as shown in Table 15. The s_{dri} symbol is then multiplied with the pilot symbol $p(k)$ before performing the spatial processing. The second symbol of steered reference for the RACH may be expressed as:

$$\underline{\mathbf{x}}(k) = \hat{\underline{\mathbf{K}}}_{\mathrm{ut}}(k) \cdot \hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k) \cdot s_{dri} \cdot p(k) \quad \text{, for } k \in K'. \qquad \text{Eq (65)}$$

As shown in equations (64) and (65), only eigenvector $\hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k)$ for the principal eigenmode is used for the steered reference for the RACH.

[00394] A symbol of steered reference transmitted by the user terminal in the preamble of the RCH may be expressed as:

$$\underline{\mathbf{x}}_{\mathrm{up,sr},m}(k) = \hat{\underline{\mathbf{K}}}_{\mathrm{ut}}(k) \cdot \hat{\underline{\mathbf{v}}}_{\mathrm{ut},m}(k) \cdot p(k) \quad \text{, for } k \in K' , \qquad \text{Eq (66)}$$

where $\underline{\mathbf{x}}_{\mathrm{up,sr},m}(k)$ is the transmit vector for subband k of wideband eigenmode m; and

$\hat{\underline{\mathbf{v}}}_{\mathrm{ut},m}(k)$ is the steering vector for subband k of wideband eigenmode m (i.e., the m-th column of $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$).

[00395] A symbol of steered reference transmitted by the access point in the preamble of the FCH may be expressed as:

$$\underline{\mathbf{x}}_{\mathrm{dn,sr},m}(k) = \hat{\underline{\mathbf{K}}}_{\mathrm{ap}}(k) \cdot \hat{\underline{\mathbf{u}}}^{*}_{\mathrm{ap},m}(k) p(k) \quad \text{, for } k \in K' , \qquad \text{Eq (67)}$$

where $\underline{\mathbf{x}}_{\mathrm{dn,sr},m}(k)$ is the transmit vector for subband k of wideband eigenmode m;

$\hat{\underline{\mathbf{K}}}_{\mathrm{ap}}(k)$ is the correction matrix for subband k for the access point; and

$\hat{\underline{\mathbf{u}}}^{*}_{\mathrm{ap},m}(k)$ is the steering vector for subband k of wideband eigenmode m.

The steering vector $\hat{\underline{\mathbf{u}}}_{\mathrm{ap},m}(k)$ is the m-th column of the matrix $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)$ of right eigenvectors of the calibrated downlink channel response estimate $\hat{\underline{\mathbf{H}}}_{\mathrm{cdn}}(k)$, where $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k) = [\hat{\underline{\mathbf{u}}}_{\mathrm{ap},1}(k)\ \hat{\underline{\mathbf{u}}}_{\mathrm{ap},2}(k)\ \hat{\underline{\mathbf{u}}}_{\mathrm{ap},3}(k)\ \hat{\underline{\mathbf{u}}}_{\mathrm{ap},4}(k)]$.

[00396] The steered reference may be transmitted in various manners. In one embodiment, one or more eigenvectors are used for the steered reference for each TDD frame and are dependent on the duration of the steered reference, which is indicated by the FCH/RCH Preamble Type fields in the FCCH information element. Table 36 lists the eigenmodes used for the preamble for the FCH and RCH for various preamble sizes, for an exemplary design.

Table 36

Type	Preamble Size	Eigenmodes Used
0	0 OFDM symbol	no preamble
1	1 OFDM symbol	eigenmode m, where m = frame counter mod 4
2	4 OFDM symbols	cycle through all 4 eigenmodes in the preamble

3	8 OFDM symbols	cycle through all 4 eigenmodes twice in the preamble

[00397] As shown in Table 36, the steered reference is transmitted for all four eigenmodes within a single TDD frame when the preamble size is four or eight OFDM symbols. The steered reference transmitted by the user terminal for the n-th OFDM symbol in the preamble for the RCH may be expressed as:

$$\underline{\mathbf{x}}_{\mathrm{up,sr},n}(k) = \underline{\hat{\mathbf{K}}}_{\mathrm{ut}}(k) \cdot \underline{\hat{\mathbf{v}}}_{\mathrm{ut},n \bmod 4}(k) \cdot p(k) \quad \text{, for } k \in K' \text{ and } n = \{1, ..., \mathrm{L}\}, \qquad \text{Eq (68)}$$

where L is the preamble size, i.e., $\mathrm{L} = 4$ for Type 2 and $\mathrm{L} = 8$ for Type 3.

[00398] Similarly, the steered reference transmitted by the access point for the n-th OFDM symbol in the preamble for the FCH may be expressed as:

$$\underline{\mathbf{x}}_{\mathrm{dn,sr},n}(k) = \underline{\hat{\mathbf{K}}}_{\mathrm{ap}}(k) \cdot \underline{\hat{\mathbf{u}}}^{*}_{\mathrm{ap},n \bmod 4}(k) p(k) \quad \text{, for } k \in K' \text{ and } n = \{1, ..., \mathrm{L}\}. \qquad \text{Eq (69)}$$

As shown in equations (68) and (69), the four eigenmodes are cycled through in each 4-symbol period by the (n mod 4) operation for the steering vector. This scheme may be used if the channel changes more rapidly and/or during the early part of a connection when a good channel estimate needs to be obtained quickly for proper system operation.

[00399] In another embodiment, the steered reference is transmitted for one wideband eigenmode for each TDD frame. The steered reference for four wideband eigenmodes may be cycled through in four TDD frames. For example, the steering vectors $\underline{\hat{\mathbf{v}}}_{\mathrm{ut},1}(k)$, $\underline{\hat{\mathbf{v}}}_{\mathrm{ut},2}(k)$, $\underline{\hat{\mathbf{v}}}_{\mathrm{ut},3}(k)$, and $\underline{\hat{\mathbf{v}}}_{\mathrm{ut},4}(k)$ may be used for the first, second, third, and fourth TDD frames, respectively, by the user terminal. The particular steering vector to use may be specified by the 2 LSBs of the Frame Counter value in the BCH message. This scheme allows a shorter preamble portion to be used in the PDU but may require a longer time period to obtain a good estimate of the channel.

[00400] For both embodiments described above, the steered reference may be transmitted on all four eigenmodes that may be used for data transmission, even though fewer than four eigenmodes are currently used (e.g., because the unused eigenmodes are poor and discarded by the water-filling). The transmission of the steered reference on a currently unused eigenmode allows the receiver to determine when the eigenmode improves enough to be selected for use.

B. <u>Steered Reference for Beam-Steering</u>

[00401] For the beam-steering mode, the spatial processing on the transmit side is performed using a set of normalized eigenvectors for the principal wideband

eigenmode. The overall transfer function with a normalized eigenvector is different from the overall transfer function with an unnormalized eigenvector (i.e., $\underline{\mathbf{H}}_{\text{cup}}(k)\hat{\underline{\mathbf{v}}}_{\text{ut},1}(k) \neq \underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)$). A steered reference generated using the set of normalized eigenvectors for all subbands may then be sent by the transmitter and used by the receiver to derive the matched filter vectors for these subbands for the beam-steering mode.

[00402] For the uplink, the steered reference for the beam-steering mode may be expressed as:

$$\tilde{\underline{\mathbf{x}}}_{\text{up,sr}}(k) = \hat{\underline{\mathbf{K}}}_{\text{ut}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)p(k) \quad \text{, for } k \in K . \qquad \text{Eq (70)}$$

At the access point, the receive uplink steered reference for the beam-steering mode may be expressed as:

$$\begin{aligned} \tilde{\underline{\mathbf{r}}}_{\text{up,sr}}(k) &= \underline{\mathbf{H}}_{\text{up}}(k)\underline{\mathbf{x}}_{\text{up,sr}}(k) + \underline{\mathbf{n}}_{\text{up}}(k) \qquad \text{, for } k \in K . \qquad \text{Eq (71)} \\ &= \underline{\mathbf{H}}_{\text{up}}(k)\hat{\underline{\mathbf{K}}}_{\text{ut}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)p(k) + \underline{\mathbf{n}}_{\text{up}}(k) \\ &= \underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)p(k) + \underline{\mathbf{n}}_{\text{up}}(k) \end{aligned}$$

[00403] To obtain the matched filter row vector $\tilde{\underline{\mathbf{m}}}_{\text{ap}}(k)$ for the uplink transmission with beam-steering, the received vector $\tilde{\underline{\mathbf{r}}}_{\text{up,sr}}(k)$ for the steered reference is first multiplied with $p^*(k)$. The result is then integrated over multiple received steered reference symbols to form an estimate of $\underline{\mathbf{H}}_{\text{cup}}(k)\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)$. The vector $\tilde{\underline{\mathbf{m}}}_{\text{ap}}(k)$ is then the conjugate transpose of this estimate.

[00404] While operating in the beam-steering mode, the user terminal may transmit multiple symbols of steered reference, for example, one or more symbols using the normalized eigenvector $\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)$, one or more symbols using the eigenvector $\hat{\underline{\mathbf{v}}}_{\text{ut},1}(k)$ for the principal wideband eigenmode, and possibly one or more symbols using the eigenvectors for the other wideband eigenmodes. The steered reference symbols generated with $\tilde{\underline{\mathbf{v}}}_{\text{ut}}(k)$ may be used by the access point to derive the matched filter vector $\tilde{\underline{\mathbf{m}}}_{\text{ap}}(k)$. The steered reference symbols generated with $\hat{\underline{\mathbf{v}}}_{\text{ut},1}(k)$ may be used to obtain $\hat{\underline{\mathbf{u}}}_{\text{ap},1}(k)$, which may then be used to derive the normalized eigenvector $\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$ that is used for beam-steering on the downlink. The steered reference symbols generated with the eigenvectors $\hat{\underline{\mathbf{v}}}_{\text{ut},2}(k)$ through $\hat{\underline{\mathbf{v}}}_{\text{ut},N_S}(k)$ for the other eigenmodes

may be used by the access point to obtain $\hat{\underline{\mathbf{u}}}_{\text{ap},2}(k)$ through $\hat{\underline{\mathbf{u}}}_{\text{ap},N_S}(k)$ and the singular values for these other eigenmodes. This information may then be used by the access point to determine whether to use the spatial multiplexing mode or the beam-steering mode for data transmission.

[00405] For the downlink, the user terminal may derive the matched filter vector $\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)$ for the beam-steering mode based on the calibrated downlink channel response estimate $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$. In particular, the user terminal has $\hat{\underline{\mathbf{u}}}_{\text{ap},1}(k)$ from the singular value decomposition of $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$ and can derive the normalized eigenvector $\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$. The user terminal can then multiply $\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$ with $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$ to obtain $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$, and may then derive $\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)$ based on $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$. Alternatively, a steered reference may be sent by the access point using the normalized eigenvector $\tilde{\underline{\mathbf{u}}}_{\text{ap}}(k)$, and this steered reference may be processed by the user terminal in the manner described above to obtain $\tilde{\underline{\mathbf{m}}}_{\text{ut}}(k)$.

4. <u>Carrier Pilot - Uplink</u>

[00406] The OFDM subband structure described herein includes four pilot subbands with indices of -21, -7, 7, and 21. In an embodiment, a carrier pilot is transmitted on the four pilot subbands in all OFDM symbols that are not part of a preamble. The carrier pilot may be used by the receiver to track phase changes due to drifts in the oscillators at both the transmitter and receiver. This may provide improved data demodulation performance.

[00407] The carrier pilot comprises four pilot sequences, $P_{c1}(n)$, $P_{c2}(n)$, $P_{c3}(n)$, and $P_{c4}(n)$, which are transmitted on the four pilot subbands. The pilot sequences may be defined as:

$$P_{c1}(n) = P_{c2}(n) = P_{c3}(n) = -P_{c4}(n) \quad \text{, for } n = \{1,\ 2,\ \dots\ 127\}, \qquad \text{Eq (72)}$$

where n is an index for OFDM symbol period.

[00408] The pilot sequences may be defined based on various data sequences. In an embodiment, the pilot sequence $P_{c1}(n)$ is generated based on a polynomial $G(x) = x^7 + x^4 + x$, where the initial state is set to all ones and the output bits are mapped to signal values as follows: $1 \Rightarrow -1$ and $0 \Rightarrow 1$. The pilot sequence $P_{c1}(n)$, for $n = \{1,\ 2,\ \dots\ 127\}$, may then be expressed as:

$P_{c1}(n)=$ {1,1,1,1,-1,-1,-1,1,-1,-1,-1,-1,1,1,-1,1,-1,-1,1,1,-1,1,1,-1,1,1,1,1,1,1,-1,1,
1,1,-1,1,1,-1,-1,1,1,1,-1,1,-1,-1,-1,1,-1,1,-1,-1,1,-1,-1,1,1,1,1,1,-1,-1,1,1,
-1,-1,1,-1,1,-1,1,1,-1,-1,-1,1,1,-1,-1,-1,-1,1,-1,-1,1,-1,1,1,1,1,-1,1,-1,1,-1,1,
-1,-1,-1,-1,-1,1,-1,1,1,-1,1,-1,1,1,1,-1,-1,1,-1,-1,-1,1,1,1,-1,-1,-1,-1,-1,-1,-1}.

The values of "1" and "-1" in the pilot sequence $P_{c1}(n)$ may be mapped to pilot symbols using a particular modulation scheme. For example, using BPSK, a "1" may be mapped to $1+j$ and a "-1" may be mapped to $-(1+j)$. If there are more than 127 OFDM symbols, then the pilot sequence may be repeated so that $P_{c1}(n) = P_{c1}(n \bmod 127)$ for $n > 127$.

[00409] In one embodiment, the four pilot sequences are reset for each transport channel. Thus, on the downlink, the pilot sequences are reset for the first OFDM symbol of the BCH message, reset again for the first OFDM symbol of the FCCH message, and reset for the first OFDM symbol sent on the FCH. In another embodiment, the pilot sequences are reset at the start of each TDD frame and repeat as often as needed. For this embodiment, the pilot sequences may be stalled during the preamble portions of the BCH and FCH.

[00410] In the diversity mode, the four pilot sequences are mapped to four subband/antenna pairing as shown in Table 29. In particular, $P_{c1}(n)$ is used for subband -21 of antenna 1, $P_{c2}(n)$ is used for subband -7 of antenna 2, $P_{c3}(n)$ is used for subband 7 of antenna 3, and $P_{c4}(n)$ is used for subband 21 of antenna 4. Each pilot sequence is then transmitted on the associated subband and antenna.

[00411] In the spatial multiplexing mode, the four pilot sequences are transmitted on the principal eigenmode of their respective subbands. The spatial processing for the carrier pilot symbols is similar to that performed for the modulation symbols, as described above. In the beam-steering mode, the four pilot sequences are transmitted on their respective subbands using beam-steering. The beam-steering for the carrier pilot symbols is also similar to that performed for the modulation symbols.

[00412] A specific pilot structure has been described above for the MIMO WLAN system. Other pilot structures may also be used for the system, and this is within the scope of the invention.

IX. System Operation

[00413] **FIG. 12A** shows a specific embodiment of a state diagram 1200 for the operation of a user terminal. This state diagram includes four states - an *Init* state 1210, a *Dormant* state 1220, an *Access* state 1230, and a *Connected* state 1240. Each of states 1210, 1220, 1230, and 1240 may be associated with a number of substates (not shown in FIG. 12A for simplicity).

[00414] In the *Init* state, the user terminal acquires the system frequency and timing and obtains system parameters sent on the BCH. In the *Init* state, the user terminal may perform the following functions:

- System determination – the user terminal determines which carrier frequency to acquire the system on.
- Frequency/timing acquisition – the user terminal acquires the beacon pilot and adjusts its frequency and timing accordingly.
- Parameter acquisition – the user terminal processes the BCH to obtain the system parameters associated with the access point from which the downlink signal is received.

Upon completing the required functions for the *Init* state, the user terminal transitions to the *Dormant* state.

[00415] In the *Dormant* state, the user terminal periodically monitors the BCH for updated system parameters, indications of pages and broadcast messages being sent on the downlink, and so on. No radio resources are allocated to the user terminal in this state. In the *Dormant* state, the user terminal may perform the following functions:

- If a registration is warranted, the user terminal enters the *Access* state with a registration request.
- If calibration of the transmitter/receiver is warranted, the user terminal enters the *Access* state with a calibration request.
- The user terminal monitors the BCH for indication of pages and broadcast messages sent on the FCH.
- If the user terminal has data to send on the uplink, it enters the *Access* state with a resource request.
- The user terminal performs maintenance procedures such as updating the system parameters and tracking the channel.

- The user terminal may enter a slotted mode of operation for power savings, if this mode is supported by the user terminal.4

If the user terminal desires radio resources from the access point for any task, it transitions to the *Access* state. For example, the user terminal may transition to the *Access* state in response to a page or DST indicator being sent in the BCH message, for registration or request for calibration, or to request dedicated resources.

[00416] In the *Access* state, the user terminal is in the process of accessing the system. The user terminal may send short messages and/or requests for FCH/RCH resources using the RACH. The operation on the RACH is described in further detail below. If the user terminal is released by the access point, then it transitions back to the *Dormant* state. If the user terminal is assigned resources for the downlink and/or uplink, then it transitions to the *Connected* state.

[00417] In the *Connected* state, the user terminal is assigned the FCH/RCH resources, although not necessarily for every TDD frame. The user terminal may actively use the allocated resources or may be idle (while still maintaining the connection) in the *Connected* state. The user terminal remains in the *Connected* state until it is released by the access point or if it times out after no activity for a particular timeout period, in which case it transitions back to the *Dormant* state.

[00418] While in the *Dormant*, *Access*, or *Connected* state, the user terminal transitions back to the *Init* state if it is powered down or if the connection is dropped.

[00419] **FIG. 12B** shows a specific embodiment of a state diagram for *Connected* state 1240. In this embodiment, the *Connected* state includes three substates - a *Setup* substate 1260, an *Open* substate 1270, and an *Idle* substate 1280. The user terminal enters the *Setup* substate upon receiving an assignment on the FCCH.

[00420] In the *Setup* substate, the user terminal is in the process of setting up the connection and is not yet exchanging data. The connection setup may include channel estimation for the access point, rate determination, service negotiation, and so on. Upon entering the *Setup* substate, the user terminal sets a timer for a specified amount of time. If the timer expires before the user terminal leaves this substate, then it transitions back to the *Dormant* state. The user terminal transitions to the *Open* substate upon completion of the connection setup.

[00421] In the *Open* substate, the user terminal and access point exchange data on the downlink and/or uplink. While in the *Open* substate, the user terminal monitors the

BCH for system parameters and indication of page/broadcast messages. If a BCH message cannot be decoded correctly within a specified number of TDD frames, then the user terminal transitions back to the *Init* state.

[00422] The user terminal also monitors the FCCH for channel assignment, rate control, RCH timing control, and power control information. The user terminal estimates the received SNR using the BCH beacon pilot and the FCH preamble and determines the maximum rate that can be sustained reliably on the FCH.

[00423] The FCH and RCH assignments for the user terminal for each TDD frame are given by the information elements in the FCCH PDU transmitted in the current (or possibly prior) TDD frame. The user terminal may not be assigned for data transmission on the FCH and/or RCH for any given TDD frame. For each TDD frame in which the user terminal is not scheduled for data transmission, it does not receive an FCH PDU on the downlink and does not transmit on the uplink.

[00424] For each TDD frame in which the user terminal is scheduled, the data transmissions on the downlink and/or uplink are performed using the rate, transmission mode, and RCH timing offset (for the uplink) indicated in the FCCH assignments (i.e., the FCCH information elements addressed to the user terminal). The user terminal receives, demodulates, and decodes FCH PDUs sent to it. The user terminal also transmits RCH PDUs, which include the preamble and FCH data rate indicator. The user terminal adjusts the rate used on the RCH according to the rate control information contained in the FCCH assignment. If power control is being applied for the uplink transmission, then the user terminal adjusts its transmit power based on the power control commands included in the FCCH assignment. The data exchange may be bursty, in which case the user terminal may enter into the *Idle* substate whenever no data is being exchanged. The user terminal enter the *Idle* substate when directed by the access point. If the access point does not assign the FCH or RCH to the user terminal within a specified number of TDD frames, then the user terminal transitions back to the *Dormant* state and retains its MAC ID.

[00425] In the *Idle* substate, both the uplink and downlink are idling. Data is not being sent in either direction. However, the links are maintained with the steered reference and control messages. In this substate, the access point periodically assigns Idle PDUs to the user terminal on the RCH and possibly the FCH (not necessarily simultaneously). The user terminal may be able to remain in the *Connected* state indefinitely, provided

that the access point periodically assigns Idle PDUs on the FCH and RCH to maintain the link.

[00426] While in the *Idle* substate, the user terminal monitors the BCH. If a BCH message is not decoded correctly within a specified number of TDD frames, then the user terminal transitions back to the *Init* state. The user terminal also monitors the FCCH for channel assignment, rate control, RCH timing control, and power control information. The user terminal may also estimate the receive SNR and determine the maximum rate supported by the FCH. The user terminal transmits an Idle PDU on the RCH, when assigned, and sets the RCH Request bit in the Idle PDU if it has data to send. If the access point does not assign an FCH or RCH to the user terminal within a specified number of TDD frames, then the user terminal transitions back to the *Dormant* state and retains its MAC ID.

[00427] A time-out timer may be set to a particular value upon entering any of the three substates. This timer would then count down if there is no activity while in the substate. While in the *Setup*, *Active*, or *Idle* substate, the terminal would transition back to the *Dormant* state if the time-out timer expires and to the *Init* state if the connection is dropped. While in the *Active* or *Idle* substate, the terminal would also transition back to the *Dormant* state if the connection is released.

[00428] FIGS. 12A and 12B show a specific embodiment of a state diagram that may be used for the user terminal. Various other state diagrams with fewer, additional, and/or different states and substates may also be used for the system, and this is within the scope of the invention.

X. Random Access

[00429] In an embodiment, a random access scheme is employed to allow the user terminals to access the MIMO WLAN system. In an embodiment, the random access scheme is based on a slotted Aloha scheme whereby a user terminal transmits in a randomly selected RACH slot to attempt to gain access to the system. The user terminal may send multiple transmissions on the RACH until access is gained or the maximum number of access attempts has been reached. Various parameters for each RACH transmission may be changed to improve the likelihood of success, as described below.

[00430] **FIG. 13** illustrates a timeline for the RACH, which is partitioned into RACH slots. The number of RACH slots available for use in each TDD frame and the duration of the RACH slot are configurable parameters. A maximum of 32 RACH slots may be

available for use in each TDD frame. The guard interval between the end of the last RACH slot and the start of the BCH PDU for the next TDD frame is also a configurable parameter. These three parameters for the RACH can change from frame to frame and are indicated by the RACH Length, RACH Slot Size, and RACH Guard Interval fields of the BCH message.

[00431] When a user terminal desires to access the system, it first processes the BCH to obtain pertinent system parameters. The user terminal then sends a RACH PDU on the RACH. This RACH PDU includes a RACH message that contains information needed by the access point to process the access request from the user terminal. For example, the RACH message includes the user terminal's assigned MAC ID that allows the access point to identify the user terminal. A registration MAC ID (i.e., a specific MAC ID value) may be reserved for unregistered user terminals. In this case, the user terminal's long ID may be included in the Payload field of the RACH message along with the registration MAC ID.

[00432] As described above, the RACH PDU may be transmitted at one of four data rates, which are listed in Table 15. The selected rate is embedded in the preamble of the RACH PDU (as shown in FIG. 5C). The RACH PDU also has a variable length of 1, 2, 4, or 8 OFDM symbols (as also listed in Table 15), which is indicated in the Message Duration field of the RACH message.

[00433] To transmit the RACH PDU, the user terminal first determines the number of RACH slots that may be used for transmission (i.e., the number of "usable" RACH slots). This determination is made based on (1) the number of RACH slots available in the current TDD frame, (2) the duration of each RACH slot, (3) the guard interval, and (4) the length of the RACH PDU to be transmitted. The RACH PDU cannot extend beyond the end of the RACH segment of the TDD frame. Thus, if the RACH PDU is longer than one RACH slot plus the guard interval, then this PDU may not be transmitted in one or more later available RACH slots. The number of RACH slots that may be used to transmit the RACH PDU may be fewer than the number of available RACH slots, based on the factors enumerated above. The RACH segment includes a guard interval, which is provided to prevent the uplink transmission from the user terminals from interfering with the next BCH segment, which is a possibility for user terminals that do not compensate for their round trip delay.

[00434] The user terminal then randomly selects one of the usable RACH slots to transmit the RACH PDU. The user terminal then transmits the RACH PDU starting in the selected RACH slot. If the user terminal knows the round trip delay to the access point, then it can account for this delay by adjusting its timing accordingly.

[00435] When the access point receives a RACH PDU, it checks the received RACH message using the CRC included in the message. The access point discards the RACH message if the CRC fails. If the CRC passes, the access point sets the RACH Acknowledgment bit on the BCH in the subsequent TDD frame and transmits an RACH acknowledgement on the FCCH within 2 TDD frames. There may be a delay between the setting of the Acknowledgment bit on the BCH and the sending of the acknowledgment on the FCCH, which may be used to account for scheduling delay and so on. For example, if the access point receives the message on the RACH, it can set the Acknowledgment bit on the BCH and have a delay response on the FCCH. The Acknowledgment bit prevents user terminals from retrying and allows unsuccessful user terminals to retry quickly, except during busy RACH periods.

[00436] If the user terminal is performing a registration, then it uses the registration MAC ID (e.g., 0x0001). The access point responds by sending a MAC ID Assignment Message on the FCH. All other RACH transmission types include the user terminal's MAC ID assigned by the system. The access point explicitly acknowledges all correctly received RACH messages by sending acknowledgments on the FCCH using the MAC ID assigned to the user terminal.

[00437] After the user terminal sends the RACH PDU, it monitors the BCH and FCCH to determine whether or not its RACH PDU has been received and processed by the access point. The user terminal monitors the BCH to determine whether or not the RACH Acknowledgment Bit in the BCH message is set. If this bit is set, which indicates that an acknowledgment for this and/or some other user terminals is being sent on the FCCH, then the user terminal further processes the FCCH to obtain IE Type 3 information elements containing acknowledgements. Otherwise, if the RACH Acknowledgment Bit is not set, then the user terminal continues to monitor the BCH or resumes its access procedure on the RACH.

[00438] The FCCH IE Type 3 is used to carry quick acknowledgements for successful access attempts. Each acknowledgement information element contains the MAC ID associated with the user terminal for which the acknowledgment is sent. A quick

acknowledgement is used to inform the user terminal that its access request has been received but is not associated with an assignment of FCH/RCH resources. Conversely, an assignment-based acknowledgement is associated with an FCH/RCH assignment. If the user terminal receives a quick acknowledgement on the FCCH, it transitions to the *Dormant* state. If the user terminal receives an assignment-based acknowledgement, it obtains scheduling information sent along with the acknowledgment and begins using the FCH/RCH as assigned in the current TDD frame.

[00439] The user terminal resumes the access procedure on the RACH if it does not receive an acknowledgement on the FCCH within a specified number of TDD frames after transmitting the RACH PDU. In this case, the user terminal can assume that the access point did not receive the RACH PDU correctly. A counter is maintained by the user terminal to count the number of access attempts. This counter may be initialized to zero for the first access attempt and is incremented by one for each subsequent access attempt. The user terminal would terminate the access procedure if the counter value reaches the maximum number of attempts.

[00440] For each subsequent access attempt, the user terminal first determines various parameters for this access attempt including (1) the amount of time to wait before transmitting the RACH PDU, (2) the RACH slot to use for the RACH PDU transmission, and (3) the rate for the RACH PDU. To determine the amount of time to wait, the user terminal first determines the maximum amount of time to wait for the next access attempt, which is referred to as the contention window (CW). In an embodiment, the contention window (which is given in units of TDD frames) exponentially increases for each access attempt (i.e., $CW = 2^{access\_attempt}$). The contention window may also be determined based on some other function (e.g., a linear function) of the number of access attempts. The amount of time to wait for the next access attempt is then randomly selected between zero and CW. The user terminal would wait this amount of time before transmitting the RACH PDU for the next access attempt.

[00441] For the next access attempt, the user terminal reduces the rate for the RACH PDU, if the lowest rate was not used for the last access attempt. The initial rate used for the first access attempt may be selected based on the received SNR of the pilot sent on the BCH. The failure to receive an acknowledgment may be caused by the access point's failure to correctly receive the RACH PDU. Thus, the rate for the RACH PDU

in the next access attempt is reduced to improve the likelihood of correct reception by the access point.

[00442] After waiting the randomly selected wait time, the user terminal again randomly selects an RACH slot for transmission of the RACH PDU. The selection of the RACH slot for this access attempt may be performed in similar manner as that described above for the first access attempt, except that the RACH parameters (i.e., number of RACH slots, slot duration, and guard interval) for the current TDD frame, as conveyed in the BCH message, are used along with the current RACH PDU length. The RACH PDU is then transmitted in the randomly selected RACH slot.

[00443] The access procedure described above continues until either (1) the user terminal receives an acknowledgment from the access point or (2) the maximum number of permitted access attempts has been reached. For each access attempt, the amount of time to wait before transmitting the RACH PDU, the RACH slot to use for the RACH PDU transmission, and the rate for the RACH PDU may be selected as described above. If the acknowledgment is received, then the user terminal operates as indicated by the acknowledgment (i.e., it waits in the *Dormant* state if a quick acknowledgment is received or starts using the FCH/RCH if an assignment-based acknowledgment is received). If the maximum number of permitted access attempts has been reached, then the user terminal transitions back to the *Init* state.

XI. Rate, Power, and Timing Control

[00444] The access point schedules downlink and uplink transmissions on the FCH and RCH and further controls the rates for all active user terminals. Moreover, the access point adjusts the transmit power of certain active user terminals on the uplink. Various control loops may be maintained to adjust the rate, transmit power, and timing for each active user terminal.

1. Fixed and Variable Rate Services

[00445] The access point can support both fixed and variable rate services on the FCH and RCH. Fixed-rate services may be used for voice, video, and so on. Variable rate services may be used for packet data (e.g., Web browsing).

[00446] For fixed rate services on the FCH/RCH, a fixed rate is used for the entire connection. Best effort delivery is used for the FCH and RCH (i.e., no retransmission). The access point schedules a constant number of FCH/RCH PDUs per specified time interval to satisfy the QoS requirements of the service. Depending on the delay

requirements, the access point may not need to schedule an FCH/RCH PDU for every TDD frame. Power control is exercised on the RCH but not the FCH for fixed rate services.

[00447] For variable rate services on the FCH/RCH, the rate used for the FCH/RCH is allowed to change with channel conditions. For some isochronous services (e.g., video, audio), the QoS requirements may impose a minimum rate constraint. For these services, the scheduler at the access point adjusts the FCH/RCH allocation so that a constant rate is provided. For asynchronous data services (e.g., web browsing, file transfer, and so on), a best effort delivery is provided with the option of retransmissions. For these services, the rate is the maximum that can be reliably sustained by the channel conditions. The scheduling of the FCH/RCH PDUs for the user terminals is typically a function of their QoS requirements. Whenever there's no data to send on the downlink/uplink, an Idle PDU is sent on the FCH/RCH to maintain the link. Closed loop power control is not exercised on the FCH or RCH for variable rate services.

2. Rate Control

[00448] Rate control may be used for variable rate services operating on the FCH and RCH to adapt the rate of the FCH/RCH to changing channel conditions. The rates to use for the FCH and RCH may be independently controlled. Moreover, in the spatial multiplexing mode, the rate for each wideband eigenmode of each dedicated transport channel may be independently controlled. The rate control is performed by the access point based on feedback provided by each active user terminal. The scheduler within the access point schedules data transmission and determines rate assignments for the active user terminals.

[00449] The maximum rate that can be supported on either link is a function of (1) the channel response matrix for all of the data subbands, (2) the noise level observed by the receiver, (3) the quality of the channel estimate, and possibly other factors. For a TDD system, the channel may be considered to be reciprocal for the downlink and uplink (after calibration has been performed to account for any differences at the access point and user terminal). However, this reciprocal channel does not imply that the noise floors are the same at the access point and user terminal. Thus, for a given user terminal, the rates on the FCH and RCH may be independently controlled.

[00450] Closed-loop rate control may be used for data transmission on one or more spatial channels. Closed-loop rate control may be achieved with one or multiple loops.

An inner loop estimates the channel conditions and selects a suitable rate for each spatial channel used for data transmission. The channel estimation and rate selection may be performed as described above. An outer loop may be used to estimate the quality of the data transmission received on each spatial channel and to adjust the operation of the inner loop. The data transmission quality may be quantified by packet error rate (PER), decoder metrics, and so on, or a combination thereof. For example, the outer loop may adjust the SNR offset for each spatial channel to achieve the target PER for that spatial channel. The outer loop may also direct the inner loop to select a lower rate for a spatial channel if excessive packet errors are detected for the spatial channel.

Downlink Rate Control

[00451] Each active user terminal can estimate the downlink channel based on the MIMO pilot transmitted on the BCH in each TDD frame. The access point may also transmit a steered reference in an FCH PDU sent to a specific user terminal. Using the MIMO pilot on the BCH and/or the steered reference on the FCH, the user terminal can estimate the received SNR and determine the maximum rate that can be supported on the FCH. If the user terminal is operating in the spatial multiplexing mode, then the maximum rate may be determined for each wideband eigenmode. Each user terminal can send back to the access point the maximum rate supported by each wideband eigenmode (for the spatial multiplexing mode), the maximum rate supported by the principal wideband eigenmode (for the beam-steering mode), or the maximum rate supported by the MIMO channel (for the diversity mode) in the FCH Rate Indicator field of the RCH PDU. These rates may be mapped to received SNRs, which may then be used to perform the water-filling described above. Alternatively, the user terminal may send back sufficient information (e.g., the received SNRs) to allow the access point to determine the maximum rate supported by the downlink.

[00452] The determination of whether to use the diversity, beam-steering, or spatial multiplexing mode may be made based on the feedback from the user terminal. The number of wideband eigenmodes selected for use may increase as isolation between the steering vectors improves.

[00453] **FIG. 14A** illustrates a process for controlling the rate of a downlink transmission for a user terminal. A BCH PDU is transmitted in the first segment of each TDD frame and includes the beacon and MIMO pilots that can be used by the user

terminals to estimate and track the channel. A steered reference may also be sent in the preamble of an FCH PDU sent to the user terminal. The user terminal estimates the channel based on the MIMO and/or steered reference and determines the maximum rate(s) that can be supported by the downlink. One rate is provided for each wideband eigenmode if the user terminal is operating in the spatial multiplexing mode. The user terminal then sends the rate indicator for the FCH in the FCH Rate Indicator field of the RCH PDU it sends to the access point.

[00454] The scheduler uses the maximum rates that the downlink can support for each active user terminal to schedule downlink data transmission in subsequent TDD frames. The rates and other channel assignment information for the user terminal are reflected in an information element sent on the FCCH. The rate assigned to one user terminal can impact the scheduling for other user terminals. The minimum delay between the rate determination by the user terminal and its use is approximately a single TDD frame.

[00455] Using the Gram-Schmidt ordered procedure, the access point can accurately determine the maximum rates supported on the FCH directly from the RCH preamble. This can then greatly simplify rate control.

<u>Uplink Rate Control</u>

[00456] Each user terminal transmits a steered reference on the RACH during system access and on the RCH upon being assigned FCH/RCH resources. The access point can estimate the received SNR for each of the wideband eigenmodes based on the steered reference on the RCH and determine the maximum rate supported by each wideband eigenmode. Initially, the access point may not have a good estimate of the channel to permit reliable operation at or near the maximum rate supported by each wideband eigenmode. To improve reliability, the initial rate used on the FCH/RCH may be much lower than the maximum supported rate. The access point can integrate the steered reference over a number of TDD frames to obtain improved estimate of the channel. As the estimate of the channel improves, the rate may be increased.

[00457] **FIG. 14B** illustrates a process for controlling the rate of an uplink transmission for a user terminal. When scheduled for uplink transmission, the user terminal transmits an RCH PDU that includes the reference, which is used by the access point to determine the maximum rate on the uplink. The scheduler then uses the maximum rates that the uplink can support for each active user terminal to schedule uplink data transmission in subsequent TDD frames. The rates and other channel assignment information for the

user terminal are reflected in an information element sent on the FCCH. The minimum delay between the rate determination by the access point and its use is approximately a single TDD frame.

3. Power Control

[00458] Power control may be used for uplink transmissions on the RCH (instead of rate control) for fixed-rate services. For fixed-rate services, the rate is negotiated at call setup and remains fixed for the duration of the connection. Some fixed-rate services may be associated with limited mobility requirement. In an embodiment, power control is implemented for the uplink to guard against interference among the user terminals but is not used for the downlink.

[00459] A power control mechanism is used to control the uplink transmit power of each active user terminal such that the received SNR at the access point is maintained at a level that achieves the desired service quality. This level is often referred to as the target received SNR, the operating point, or the setpoint. For a mobile user terminal, the propagation loss will likely change as the user terminal moves about. The power control mechanism tracks changes in the channel so that the received SNR is maintained near the setpoint.

[00460] The power control mechanism may be implemented with two power control loops - an inner loop and an outer loop. The inner loop adjusts the transmit power of the user terminal such that the received SNR at the access point is maintained near the setpoint. The outer loop adjusts the setpoint to achieve a particular level of performance, which may be quantified by a particular frame error rate (FER) (e.g., 1% FER), packet error rate (PER), block error rate (BLER), message error rate (MER), or some other measure.

[00461] **FIG. 15** illustrates the operation of the inner power control for a user terminal. After the user terminal is assigned the FCH/RCH, the access point estimates the received SNR on the RCH and compares it to the setpoint. The initial power to be used by the user terminal may be determined at call setup and is typically near its maximum transmit power level. For each frame interval, if the received SNR exceeds the setpoint by a particular positive margin δ, then the access point can direct the user terminal to reduce its transmit power by a particular amount (e.g., 1 dB) in an FCCH information element sent to this user terminal. Conversely, if the received SNR is lower than the threshold minus the margin δ, then the access point can direct the user terminal to

increase its transmit power by the particular amount. If the received SNR is within the acceptable limits of the setpoint, then the access point will not request a change in transmit power by the user terminal. The uplink transmit power is given as the initial transmit power level plus the sum of all power adjustments received from the access point.

[00462] The initial setpoint used at the access point is set to achieve a particular level of performance. This setpoint is adjusted by the outer loop based on the FER or PER for the RCH. For example, if no frame/packet errors occur over a specified time period, then the setpoint may be reduced by a first amount (e.g., 0.1 dB). If the average FER is exceeded by the occurrence of one or more frame/packet errors, then the setpoint may be increased by a second amount (e.g., 1 dB). The setpoint, hysteresis margin, and outer loop operation are specific to the power control design used for the system.

4. Timing Control

[00463] Timing control may be advantageously used in a TDD-based frame structure where the downlink and uplink share the same frequency band in a time division duplexed manner. The user terminals may be located throughout in the system and may thus be associated with different propagation delays to the access point. In order to maximize efficiency on the uplink, the timing of the uplink transmission on the RCH and RACH from each user terminal can be adjusted to account for its propagation delay. This would then ensure that the uplink transmissions from different user terminals arrive within a particular time window at the access point and do not interfere with one another on the uplink, or with the downlink transmission.

[00464] **FIG. 16** illustrates a process for adjusting the uplink timing of a user terminal. Initially, the user terminal sends an RACH PDU on the uplink to gain access to the system. The access point derives an initial estimate of the round trip delay (RTD) associated with the user terminal. The round trip delay may be estimated based on (1) a sliding correlator used at the access point to determine the start of transmission, and (2) the slot ID included in the RACH PDU sent by the user terminal. The access point then computes an initial timing advance for the user terminal based on the initial RTD estimate. The initial timing advance is sent to the user terminal prior to its transmission on the RCH. The initial timing advance may be sent in a message on the FCH, a field of an FCCH information element, or by some other means.

[00465] The user terminal receives the initial timing advance from the access point and thereafter uses this timing advance on all subsequent uplink transmissions on both the RCH and RACH. If the user terminal is assigned FCH/RCH resources, then its timing advance can be adjusted by commands sent by the access point in the RCH Timing Adjustment field of an FCCH information element. The user terminal would thereafter adjust its uplink transmissions on the RCH based on the current timing advance, which is equal to the initial timing advance plus all timing adjustments sent by the access point to the user terminal.

[00466] Various parts of the MIMO WLAN system and various techniques described herein may be implemented by various means. For example, the processing at the access point and user terminal may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the processing may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[00467] For a software implementation, the processing may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 732 or 782 in FIG. 7) and executed by a processor (e.g., controller 730 or 780). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[00468] Headings are included herein for reference and to aid in locating certain sections. These headings are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[00469] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be

limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. A method of transmitting data in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

selecting at least one user terminal from among a plurality of user terminals for data transmission in a current scheduling interval, wherein the at least one user terminal includes a user terminal with multiple antennas;

selecting at least one rate for each of the at least one user terminal, wherein each of the at least one rate is selected from among a plurality of rates supported by the system, and wherein each of the plurality of rates is associated with a particular code rate and a particular modulation scheme;

selecting a transmission mode for each of the at least one user terminal, wherein the transmission mode for each user terminal is selected from among a plurality of transmission modes supported by the system; and

scheduling the at least one user terminal for data transmission in the current scheduling interval with the at least one rate and the transmission mode selected for each user terminal.

2. The method of claim 1, further comprising:

selecting a transmission duration for each of the at least one user terminal, and wherein the at least one user terminal is scheduled for data transmission in the current scheduling interval for the transmission duration selected for each user terminal.

3. The method of claim 1, wherein each of the at least one user terminal is scheduled for data transmission on a downlink, an uplink, or both the downlink and uplink in the current scheduling interval.

4. The method of claim 3, wherein for each user terminal scheduled for data transmission on both the downlink and uplink, the at least one rate and the transmission mode for the user terminal are selected independently for the downlink and uplink.

5. The method of claim 2, wherein for each user terminal scheduled for data transmission on both downlink and uplink, the transmission duration for the user terminal is selected independently for the downlink and uplink.

6. The method of claim 1, wherein the plurality of transmission modes include a diversity mode and a spatial multiplexing mode, the diversity mode supporting data transmission with redundancy from a plurality of transmit antennas, and the spatial multiplexing mode supporting data transmission on a plurality of spatial channels.

7. The method of claim 6, wherein the plurality of transmission modes further include a beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among the plurality of spatial channels.

8. The method of claim 6, wherein the plurality of transmission modes further include a single-input multiple-output (SIMO) mode supporting data transmission from a single transmit antenna to multiple receive antennas.

9. The method of claim 1, wherein the transmission mode selected for each user terminal is dependent on the number of antennas available at the user terminal.

10. The method of claim 1, wherein the MIMO communication system utilizes orthogonal frequency division multiplexing (OFDM).

11. The method of claim 10, further comprising:

selecting a transmission duration, in integer number of OFDM symbols, for each of the at least one user terminal, and wherein the at least one user terminal is scheduled for data transmission in the current scheduling interval for the transmission duration selected for each user terminal.

12. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to

select at least one user terminal from among a plurality of user terminals for data transmission in a current scheduling interval, wherein the at least one user terminal includes a user terminal with multiple antennas,

select at least one rate for each of the at least one user terminal, wherein each of the at least one rate is selected from among a plurality of rates supported by the system, and wherein each of the plurality of rates is associated with a particular code rate and a particular modulation scheme, and

select a transmission mode for each of the at least one user terminal, wherein the transmission mode for each user terminal is selected from among a plurality of transmission modes supported by the system; and

a scheduler operative to schedule the at least one user terminal for data transmission in the current scheduling interval with the at least one rate and the transmission mode selected for each user terminal.

13. The apparatus of claim 12, wherein the controller is further operative to select a transmission duration for each of the at least one user terminal, and wherein the at least one user terminal is scheduled for data transmission in the current scheduling interval for the transmission duration selected for each user terminal.

14. The apparatus of claim 12, wherein the plurality of transmission modes include a diversity mode and a spatial multiplexing mode, the diversity mode supporting data transmission with redundancy from a plurality of transmit antennas, and the spatial multiplexing mode supporting data transmission on a plurality of spatial channels.

15. The apparatus of claim 14, wherein the plurality of transmission modes further include a beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among the plurality of spatial channels.

16. The apparatus of claim 12, wherein the MIMO communication system utilizes orthogonal frequency division multiplexing (OFDM).

17. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for selecting at least one user terminal from among a plurality of user terminals for data transmission in a current scheduling interval, wherein the at least one user terminal includes a user terminal with multiple antennas;

means for selecting at least one rate for each of the at least one user terminal, wherein each of the at least one rate is selected from among a plurality of rates supported by the system, and wherein each of the plurality of rates is associated with a particular code rate and a particular modulation scheme;

means for selecting a transmission mode for each of the at least one user terminal, wherein the transmission mode for each user terminal is selected from among a plurality of transmission modes supported by the system; and

means for scheduling the at least one user terminal for data transmission in the current scheduling interval with the at least one rate and the transmission mode selected for each user terminal.

18. The apparatus of claim 17, further comprising:

means for selecting a transmission duration for each of the at least one user terminal, and wherein the at least one user terminal is scheduled for data transmission in the current scheduling interval for the transmission duration selected for each user terminal.

19. The apparatus of claim 17, wherein the plurality of transmission modes include a diversity mode and a spatial multiplexing mode, the diversity mode supporting data transmission with redundancy from a plurality of transmit antennas, and the spatial multiplexing mode supporting data transmission on a plurality of spatial channels.

20. The apparatus of claim 19, wherein the plurality of transmission modes further include a beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among the plurality of spatial channels.

21. The apparatus of claim 17, wherein the MIMO communication system utilizes orthogonal frequency division multiplexing (OFDM).

22. A method of transmitting data in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

selecting a first user terminal, equipped with a single receive antenna, from among a plurality of user terminals;

transmitting data, based on a first transmission mode, from multiple transmit antennas to the single received antenna of the first user terminal in a first time interval;

selecting a second user terminal, equipped with multiple receive antennas, from among the plurality of user terminals; and

transmitting data, based on a second transmission mode, from the multiple transmit antennas to the multiple receive antennas of the second user terminal in a second time interval, wherein the first and second transmission modes are selected from among a plurality of transmission modes supported by the system.

23. The method of claim 22, further comprising:

selecting a third user terminal, equipped with multiple receive antennas, from among the plurality of user terminals; and

transmitting data, based on a third transmission mode, from the multiple transmit antennas to the multiple receive antennas of the third user terminal in a third time interval, wherein the third transmission mode is selected from among the plurality of transmission modes.

24. The method of claim 22, wherein the plurality of transmission modes include a spatial multiplexing mode supporting data transmission on a plurality of spatial channels formed by the multiple transmit antennas and multiple receive antennas.

25. The method of claim 24, wherein each of the plurality of spatial channels is associated with a respective rate.

26. The method of claim 24, wherein the number of spatial channels used for data transmission in the spatial multiplexing mode is selectable.

27. The method of claim 22, wherein the plurality of transmission modes include a beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among a plurality of spatial channels formed by the multiple transmit antennas and multiple receive antennas.

28. The method of claim 22, wherein the plurality of transmission modes include a diversity mode supporting data transmission with redundancy from the multiple transmit antennas.

29. The method of claim 28, wherein the diversity mode implements space-time transmit diversity (STTD) supporting transmission of each pair of modulation symbols from a pair of antennas in two symbol periods.

30. The method of claim 28, wherein the diversity mode implements space-frequency transmit diversity (SFTD) supporting transmission of each pair of modulation symbols from a pair of antennas in two subbands.

31. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to select a first user terminal equipped with a single receive antenna and a second user terminal equipped with multiple receive antennas from among a plurality of user terminals; and

a transmit spatial processor operative to

process data based on a first transmission mode for transmission from multiple transmit antennas to the single received antenna of the first user terminal in a first time interval, and

process data based on a second transmission mode for transmission from the multiple transmit antennas to the multiple receive antennas of the second user terminal in a second time interval, wherein the first and second transmission modes are selected from among a plurality of transmission modes supported by the system.

32. The apparatus of claim 31, wherein the controller is further operative to select a third user terminal equipped with multiple receive antennas from among the plurality of user terminals, and wherein the transmit spatial processor is further operative to process data based on a third transmission mode for transmission from the multiple transmit antennas to the multiple receive antennas of the third user terminal in a third time interval, wherein the third transmission mode is selected from among the plurality of transmission modes.

33. The apparatus of claim 31, wherein the plurality of transmission modes include a spatial multiplexing mode supporting data transmission on a plurality of spatial channels formed by the multiple transmit antennas and multiple receive antennas.

34. The apparatus of claim 31, wherein the plurality of transmission modes include a beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among a plurality of spatial channels formed by the multiple transmit antennas and multiple receive antennas.

35. The apparatus of claim 31, wherein the plurality of transmission modes include a diversity mode supporting data transmission with redundancy from the multiple transmit antennas.

36. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for selecting a first user terminal, equipped with a single receive antenna, from among a plurality of user terminals;

means for transmitting data, based on a first transmission mode, from multiple transmit antennas to the single received antenna of the first user terminal in a first time interval;

means for selecting a second user terminal, equipped with multiple receive antennas, from among the plurality of user terminals; and

means for transmitting data, based on a second transmission mode, from the multiple transmit antennas to the multiple receive antennas of the second user terminal in a second time interval, wherein the first and second transmission modes are selected from among a plurality of transmission modes supported by the system.

37. The apparatus of claim 36, further comprising:

means for selecting a third user terminal, equipped with multiple receive antennas, from among the plurality of user terminals; and

means for transmitting data, based on a third transmission mode, from the multiple transmit antennas to the multiple receive antennas of the third user terminal in a third time interval, wherein the third transmission mode is selected from among the plurality of transmission modes.

38. The apparatus of claim 36, wherein the plurality of transmission modes include a spatial multiplexing mode supporting data transmission on a plurality of spatial channels formed by the multiple transmit antennas and multiple receive antennas.

39. The apparatus of claim 36, wherein the plurality of transmission modes include a beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among a plurality of spatial channels formed by the multiple transmit antennas and multiple receive antennas.

40. The apparatus of claim 36, wherein the plurality of transmission modes include a diversity mode supporting data transmission with redundancy from the multiple transmit antennas.

41. A method of exchanging data in a wireless time division duplex (TDD) multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

selecting a first set of at least one user terminal for data transmission on a downlink in a current scheduling interval;

selecting a second set of at least one user terminal for data transmission on an uplink in the current scheduling interval;

transmitting data on the downlink to the first set of at least one user terminal in a first time segment of the current scheduling interval; and

receiving data transmission on the uplink from the second set of at least one user terminal in a second time segment of the current scheduling interval, wherein the first and second time segments are time division duplexed in the current scheduling interval.

42. The method of claim 41, further comprising:

selecting a transmission mode, from among a plurality of transmission modes supported by the system, for each user terminal in the first set, and wherein data is transmitted to each user terminal in the first set is based on the transmission mode selected for the user terminal.

43. The method of claim 41, further comprising:

obtaining channel estimates for each user terminal in the first set based on a pilot transmitted on the uplink by the user terminal, and wherein data is transmitted to each user terminal in the first set based on the channel estimates obtained for the user terminal.

44. The method of claim 41, further comprising:

selecting a transmission mode, from among a plurality of transmission modes supported by the system, for each user terminal in the second set, and wherein the data transmission from each user terminal in the second set is based on the transmission mode selected for the user terminal.

45. The method of claim 41, further comprising:

determining timing of each user terminal in the second set; and

adjusting timing of the data transmission on the uplink for each user terminal in the second set based on the determined timing of the user terminal.

46. The method of claim 41, further comprising:

determining received power for each user terminal in the second set; and

adjusting transmit power of the data transmission on the uplink for each user terminal in the second set based on the received power for the user terminal.

47. An apparatus in a wireless time division duplex (TDD) multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to select a first set of at least one user terminal for data transmission on a downlink in a current scheduling interval and a second set of at least one user terminal for data transmission on an uplink in the current scheduling interval;

a transmit spatial processor operative to process data for transmission on the downlink to the first set of at least one user terminal in a first time segment of the current scheduling interval; and

a receive spatial processor operative to receive data transmission on the uplink from the second set of at least one user terminal in a second time segment of the current scheduling interval, wherein the first and second time segments are time division duplexed in the current scheduling interval.

48. The apparatus of claim 47, wherein the controller is further operative to select a transmission mode, from among a plurality of transmission modes supported by the system, for each user terminal in the first set, and wherein data is transmitted to each user terminal in the first set is based on the transmission mode selected for the user terminal.

49. The apparatus of claim 47, wherein the controller is further operative to obtain channel estimates for each user terminal in the first set based on a pilot transmitted on the uplink by the user terminal, and wherein data is transmitted to each user terminal in the first set based on the channel estimates obtained for the user terminal.

50. The apparatus of claim 47, wherein the controller is further operative to select a transmission mode, from among a plurality of transmission modes supported by the system, for each user terminal in the second set, and wherein the data transmission from each user terminal in the second set is based on the transmission mode selected for the user terminal.

51. The apparatus of claim 47, wherein the controller is further operative to determine timing of each user terminal in the second set and to adjust timing of the data transmission on the uplink for each user terminal in the second set based on the determined timing of the user terminal.

52. The apparatus of claim 47, wherein the controller is further operative to determine received power for each user terminal in the second set and to adjust transmit

power of the data transmission on the uplink for each user terminal in the second set based on the received power for the user terminal.

53. An apparatus in a wireless time division duplex (TDD) multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for selecting a first set of at least one user terminal for data transmission on a downlink in a current scheduling interval;

means for selecting a second set of at least one user terminal for data transmission on an uplink in the current scheduling interval;

means for transmitting data on the downlink to the first set of at least one user terminal in a first time segment of the current scheduling interval; and

means for receiving data transmission on the uplink from the second set of at least one user terminal in a second time segment of the current scheduling interval, wherein the first and second time segments are time division duplexed in the current scheduling interval.

54. The apparatus of claim 53, further comprising:

means for selecting a transmission mode, from among a plurality of transmission modes supported by the system, for each user terminal in the first set, and wherein data is transmitted to each user terminal in the first set is based on the transmission mode selected for the user terminal.

55. The apparatus of claim 53, further comprising:

means for obtaining channel estimates for each user terminal in the first set based on a pilot transmitted on the uplink by the user terminal, and wherein data is transmitted to each user terminal in the first set based on the channel estimates obtained for the user terminal.

56. The apparatus of claim 53, further comprising:

means for selecting a transmission mode, from among a plurality of transmission modes supported by the system, for each user terminal in the second set, and wherein the data transmission from each user terminal in the second set is based on the transmission mode selected for the user terminal.

57. The apparatus of claim 53, further comprising:

means for determining timing of each user terminal in the second set; and

means for adjusting timing of the data transmission on the uplink for each user terminal in the second set based on the determined timing of the user terminal.

58. The apparatus of claim 53, further comprising:

means for determining received power for each user terminal in the second set; and

means for adjusting transmit power of the data transmission on the uplink for each user terminal in the second set based on the received power for the user terminal.

59. A method of exchanging data in a wireless time division duplex (TDD) multiple-input multiple-output (MIMO) communication system, comprising:

receiving a pilot on an uplink from a user terminal;

deriving at least one steering vector for a downlink for the user terminal based on the received pilot; and

performing spatial processing, with the at least one steering vector, on a first data transmission sent on the downlink to the user terminal.

60. The method of claim 59, wherein a single steering vector is derived for the downlink for the user terminal, and wherein spatial processing for beam-steering is performed on the first data transmission with the single steering vector to send the first data transmission via a single spatial channel of the downlink.

61. The method of claim 59, wherein a plurality of steering vectors are derived for the downlink for the user terminal, and wherein spatial processing for spatial multiplexing is performed on first the data transmission with the plurality of steering vectors to send the first data transmission via a plurality of spatial channels of the downlink.

62. The method of claim 59, further comprising:

deriving a matched filter for the uplink for the user terminal based on the received pilot; and

performing matched filtering of a second data transmission received on the uplink from the user terminal with the matched filter.

63. The method of claim 62, wherein the matched filter comprises at least one eigenvector for at least one eigenmode of the uplink, and wherein the at least one eigenvector for the uplink is equal to the at least one steering vector for the downlink.

64. An apparatus in a wireless time division duplex (TDD) multiple-input multiple-output (MIMO) communication system, comprising:

a receive spatial processor operative to receive a pilot on an uplink from a user terminal;

a controller operative to derive at least one steering vector for a downlink for the user terminal based on the received pilot; and

a transmit spatial processor operative to perform spatial processing with the at least one steering vector on a first data transmission sent on the downlink to the user terminal.

65. The apparatus of claim 64, wherein the controller is operative to derive a single steering vector for the downlink for the user terminal, and wherein the transmit spatial processor is operative to perform spatial processing for beam-steering on the first data transmission with the single steering vector to send the first data transmission via a single spatial channel of the downlink.

66. The apparatus of claim 64, wherein the controller is operative to derive a plurality of steering vectors for the downlink for the user terminal, and wherein the transmit spatial processor is operative to perform spatial processing for spatial multiplexing on first the data transmission with the plurality of steering vectors to send the first data transmission via a plurality of spatial channels of the downlink.

67. The apparatus of claim 64, wherein the controller is further operative to derive a matched filter for the uplink for the user terminal based on the received pilot,

and wherein the receive spatial processor is further operative to perform matched filtering of a second data transmission received on the uplink from the user terminal with the matched filter.

68. The apparatus of claim 67, wherein the matched filter comprises at least one eigenvector for at least one eigenmode of the uplink, and wherein the at least one eigenvector for the uplink is equal to the at least one steering vector for the downlink.

69. An apparatus in a wireless time division duplex (TDD) multiple-input multiple-output (MIMO) communication system, comprising:

means for receiving a pilot on an uplink from a user terminal;

means for deriving at least one steering vector for a downlink for the user terminal based on the received pilot; and

means for performing spatial processing, with the at least one steering vector, on a first data transmission sent on the downlink to the user terminal.

70. The apparatus of claim 69, wherein a single steering vector is derived for the downlink for the user terminal, and wherein spatial processing for beam-steering is performed on the first data transmission with the single steering vector to send the first data transmission via a single spatial channel of the downlink.

71. The apparatus of claim 69, wherein a plurality of steering vectors are derived for the downlink for the user terminal, and wherein spatial processing for spatial multiplexing is performed on first the data transmission with the plurality of steering vectors to send the first data transmission via a plurality of spatial channels of the downlink.

72. The apparatus of claim 69, further comprising:

means for deriving a matched filter for the uplink for the user terminal based on the received pilot; and

means for performing matched filtering of a second data transmission received on the uplink from the user terminal with the matched filter.

73. The apparatus of claim 72, wherein the matched filter comprises at least one eigenvector for at least one eigenmode of the uplink, and wherein the at least one eigenvector for the uplink is equal to the at least one steering vector for the downlink.

74. A method of transmitting and receiving pilots in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

transmitting a MIMO pilot from a plurality of antennas and on a first communication link, wherein the MIMO pilot comprises a plurality of pilot transmissions sent from the plurality of antennas, and wherein the pilot transmission from each antenna is identifiable by a communicating entity receiving the MIMO pilot; and

receiving a steered pilot via at least one eigenmode of a second communication link from the communicating entity, wherein the steered pilot is generated based on the MIMO pilot.

75. The method of claim 74, wherein the first communication link is an uplink, the second communication link is a downlink, and the communicating entity is a user terminal.

76. The method of claim 74, wherein the first communication link is a downlink, the second communication link is an uplink, and the communicating entity is an access point.

77. The method of claim 74, wherein the pilot transmission from each antenna is associated with a different orthogonal code.

78. The method of claim 74, wherein the steered pilot is received via a single eigenmode of the second communication link and is transmitted at full transmit power from a plurality of antennas at the communicating entity.

79. The method of claim 74, wherein the steered pilot is received via a plurality of eigenmodes of the second communication link.

80. The method of claim 74, wherein the steered pilot is transmitted by the communicating entity for a time duration configurable by the system.

81. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

a transmit spatial processor operative to generate a MIMO pilot for transmission from a plurality of antennas and on a first communication link, wherein the MIMO pilot comprises a plurality of pilot transmissions sent from the plurality of antennas, and wherein the pilot transmission from each antenna is identifiable by a communicating entity receiving the MIMO pilot; and

a receive spatial processor operative to process a steered pilot received via at least one eigenmode of a second communication link from the communicating entity, wherein the steered pilot is generated based on the MIMO pilot.

82. The apparatus of claim 81, wherein the pilot transmission from each antenna is associated with a different orthogonal code.

83. The apparatus of claim 81, wherein the steered pilot is received via a single eigenmode of the second communication link and is transmitted at full transmit power from a plurality of antennas at the communicating entity.

84. The apparatus of claim 81, wherein the steered pilot is received via a plurality of eigenmodes of the second communication link.

85. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

means for transmitting a MIMO pilot from a plurality of antennas and on a first communication link, wherein the MIMO pilot comprises a plurality of pilot transmissions sent from the plurality of antennas, and wherein the pilot transmission from each antenna is identifiable by a communicating entity receiving the MIMO pilot; and

means for receiving a steered pilot via at least one eigenmode of a second communication link from the communicating entity, wherein the steered pilot is generated based on the MIMO pilot.

86. The apparatus of claim 85, wherein the pilot transmission from each antenna is associated with a different orthogonal code.

87. The apparatus of claim 85, wherein the steered pilot is received via a single eigenmode of the second communication link and is transmitted at full transmit power from a plurality of antennas at the communicating entity.

88. The apparatus of claim 85, wherein the steered pilot is received via a plurality of eigenmodes of the second communication link.

89. A method of performing channel estimation in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

receiving a steered pilot via at least one eigenmode of an uplink from a user terminal; and

estimating a channel response of the at least one eigenmode of the uplink for the user terminal based on the received steered pilot.

90. The method of claim 89, further comprising:

deriving a matched filter based on the estimated channel response of the at least one eigenmode of the uplink, wherein the matched filter is used for matched filtering of a data transmission received via the at least one eigenmode of the uplink from the user terminal.

91. The method of claim 89, further comprising:

estimating a channel response of at least one eigenmode of a downlink for the user terminal based on the received steered pilot.

92. The method of claim 91, further comprising:

deriving at least one steering vector based on the estimated channel response of the at least one eigenmode of the downlink, wherein the at least one steering vector is used for data transmission on the at least one eigenmode of the downlink to the user terminal.

93. The method of claim 92, wherein the steered pilot is received via a plurality of eigenmodes of the uplink, wherein the channel response of a plurality of eigenmode of the downlink for the user terminal is estimated based on the received steered pilot, and wherein a plurality of steering vectors are derived based on the estimated channel response of the plurality of eigenmodes of the downlink.

94. The method of claim 93, wherein the plurality of steering vectors are derived to be orthogonal to one another.

95. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

a receive spatial processor operative to receive a steered pilot via at least one eigenmode of an uplink from a user terminal; and

a controller operative to estimate a channel response of the at least one eigenmode of the uplink for the user terminal based on the received steered pilot.

96. The apparatus of claim 95, wherein the controller is further operative to derive a matched filter based on the estimated channel response of the at least one eigenmode of the uplink, wherein the matched filter is used for matched filtering of a data transmission received via the at least one eigenmode of the uplink from the user terminal.

97. The apparatus of claim 95, wherein the controller is further operative to estimate a channel response of at least one eigenmode of a downlink for the user terminal based on the received steered pilot.

98. The apparatus of claim 97, wherein the controller is further operative to derive at least one steering vector based on the estimated channel response of the at

least one eigenmode of the downlink, wherein the at least one steering vector is used for data transmission on the at least one eigenmode of the downlink to the user terminal.

99. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

means for receiving a steered pilot via at least one eigenmode of an uplink from a user terminal; and

means for estimating a channel response of the at least one eigenmode of the uplink for the user terminal based on the received steered pilot.

100. The apparatus of claim 99, further comprising:

means for deriving a matched filter based on the estimated channel response of the at least one eigenmode of the uplink, wherein the matched filter is used for matched filtering of a data transmission received via the at least one eigenmode of the uplink from the user terminal.

101. The apparatus of claim 99, further comprising:

means for estimating a channel response of at least one eigenmode of a downlink for the user terminal based on the received steered pilot.

102. The apparatus of claim 101, further comprising:

means for deriving at least one steering vector based on the estimated channel response of the at least one eigenmode of the downlink, wherein the at least one steering vector is used for data transmission on the at least one eigenmode of the downlink to the user terminal.

103. A channel structure for a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a broadcast channel for transmitting, on a downlink, system parameters and a pilot used for channel estimation of the downlink;

a forward control channel for transmitting, on the downlink, a schedule for data transmission on the downlink and an uplink;

a forward channel for transmitting traffic data on the downlink;

a random access channel for transmitting, on the uplink, user requests to access the system; and

a reverse channel for transmitting traffic data on the uplink.

104. The channel structure of claim 103, wherein the broadcast channel, forward control channel, forward channel, random access channel, and reverse channel are time division multiplexed within a frame having a predetermined time duration.

105. The channel structure of claim 104, wherein the broadcast channel is transmitted first and the forward control channel is transmitted second in the frame.

106. The channel structure of claim 103, wherein the broadcast channel and the forward control channel are transmitted using a diversity mode supporting data transmission with redundancy from a plurality of transmit antennas.

107. The channel structure of claim 103, wherein the forward channel and the reverse channel support a diversity mode and a spatial multiplexing mode, the diversity mode supporting data transmission with redundancy from a plurality of transmit antennas, and the spatial multiplexing mode supporting data transmission on a plurality of spatial channels.

108. The channel structure of claim 103, wherein the random access channel supports a single-input multiple-output (SIMO) mode and a beam-steering mode, the SIMO mode supporting data transmission from a single transmit antenna to multiple receive antennas, and the beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among a plurality of spatial channels.

109. The channel structure of claim 103, wherein the forward channel and the reverse channel each has a variable time duration.

110. The channel structure of claim 103, wherein the forward control channel and the random access channel each has a variable time duration.

111. The channel structure of claim 103, wherein the schedule includes identities of user terminals scheduled for data transmission on the downlink and uplink.

112. The channel structure of claim 103, wherein the schedule includes a transmission mode and at least one rate for each user terminal scheduled for data transmission on the downlink and uplink, the transmission mode being selected from among a plurality of transmission modes supported by the system, and each of the at least one rate being selected from among a plurality of rates supported by the system.

113. The channel structure of claim 103, wherein the forward channel is further for transmitting a steered pilot on at least one eigenmode of the downlink for a user terminal.

114. The channel structure of claim 103, wherein the reverse channel is further for transmitting on the uplink a second pilot used for channel estimation of the uplink.

115. The channel structure of claim 103, wherein the reverse channel is further for transmitting a steered pilot on at least one eigenmode of the uplink from a user terminal.

116. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a transmit data processor operative to

process system parameters and a pilot for transmission via a broadcast channel, wherein the pilot is used for channel estimation of the downlink,

process scheduling information for transmission via a forward control channel, wherein the scheduling information is for data transmission on the downlink and an uplink, and

process traffic data for transmission via a forward channel; and

a receive data processor operative to

process user requests received via a random access channel, and

process traffic data received via a reverse channel.

117. The apparatus of claim 116, wherein the broadcast channel, forward control channel, forward channel, random access channel, and reverse channel are time division multiplexed within a frame having a predetermined time duration.

118. The apparatus of claim 116, wherein the broadcast channel and the forward control channel are transmitted using a diversity mode supporting data transmission with redundancy from a plurality of transmit antennas.

119. The apparatus of claim 116, wherein the forward channel and the reverse channel support a diversity mode and a spatial multiplexing mode, the diversity mode supporting data transmission with redundancy from a plurality of transmit antennas, and the spatial multiplexing mode supporting data transmission on a plurality of spatial channels.

120. The apparatus of claim 116, wherein the random access channel supports a single-input multiple-output (SIMO) mode and a beam-steering mode, the SIMO mode supporting data transmission from a single transmit antenna to multiple receive antennas, and the beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among a plurality of spatial channels.

121. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for processing system parameters and a pilot for transmission via a broadcast channel, wherein the pilot is used for channel estimation of the downlink;

means for processing scheduling information for transmission via a forward control channel, wherein the scheduling information is for data transmission on the downlink and an uplink;

means for processing traffic data for transmission via a forward channel;

means for processing user requests received via a random access channel; and

means for processing traffic data received via a reverse channel.

122. The channel structure of claim 121, wherein the broadcast channel, forward control channel, forward channel, random access channel, and reverse channel are time division multiplexed within a frame having a predetermined time duration.

123. The channel structure of claim 121, wherein the broadcast channel and the forward control channel are transmitted using a diversity mode supporting data transmission with redundancy from a plurality of transmit antennas.

124. The channel structure of claim 121, wherein the forward channel and the reverse channel support a diversity mode and a spatial multiplexing mode, the diversity mode supporting data transmission with redundancy from a plurality of transmit antennas, and the spatial multiplexing mode supporting data transmission on a plurality of spatial channels.

125. The channel structure of claim 121, wherein the random access channel supports a single-input multiple-output (SIMO) mode and a beam-steering mode, the SIMO mode supporting data transmission from a single transmit antenna to multiple receive antennas, and the beam-steering mode supporting data transmission on a single spatial channel associated with a highest rate among a plurality of spatial channels.

126. A method of transmitting signaling information in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

transmitting signaling information for a first set of at least one user terminal at a first rate on a first subchannel of a forward control channel; and

transmitting signaling information for a second set of at least one user terminal at a second rate on a second subchannel of the forward control channel, wherein the second rate is higher than the first rate, and wherein the second subchannel is transmitted after the first subchannel.

127. The method of claim 126, further comprising:

transmitting signaling information for a third set of at least one user terminal at a third rate on a third subchannel of the forward control channel, wherein the third rate is

higher than the second rate, and wherein the third subchannel is transmitted after the second subchannel.

128. The method of claim 126, wherein the first subchannel indicates whether or not the second subchannel is transmitted in a current frame.

129. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

a transmit data processor operative to

process signaling information for a first set of at least one user terminal based on a first rate, and

process signaling information for a second set of at least one user terminal based on a second rate that is higher than the first rate; and

a transmitter unit operative to

transmit the processed scheduling information for the first user terminal set on a first subchannel of a forward control channel, and

transmit the processed scheduling information for the second user terminal set on a second subchannel of the forward control channel, wherein the second subchannel is transmitted after the first subchannel.

130. The apparatus of claim 129, wherein the transmit data processor is further operative to process signaling information for a third set of at least one user terminal based on a third rate that is higher than the second rate, and wherein the transmitter unit is further operative to transmit the processed signaling information for the third user terminal set on a third subchannel of the forward control channel, wherein the third subchannel is transmitted after the second subchannel.

131. The apparatus of claim 129, wherein the first subchannel indicates whether or not the second subchannel is transmitted in a current frame.

132. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

means for transmitting signaling information for a first set of at least one user terminal at a first rate on a first subchannel of a forward control channel; and

means for transmitting signaling information for a second set of at least one user terminal at a second rate on a second subchannel of the forward control channel, wherein the second rate is higher than the first rate, and wherein the second subchannel is transmitted after the first subchannel.

133. The apparatus of claim 132, further comprising:

means for transmitting signaling information for a third set of at least one user terminal at a third rate on a third subchannel of the forward control channel, wherein the third rate is higher than the second rate, and wherein the third subchannel is transmitted after the second subchannel.

134. The apparatus of claim 132, wherein the first subchannel indicates whether or not the second subchannel is transmitted in a current frame.

135. A method of receiving signaling information at a user terminal in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

receiving signaling information sent at a first rate on a first subchannel of a forward control channel; and

if signaling information for the user terminal is not obtained from the first subchannel, receiving signaling information sent at a second rate on a second subchannel of the forward control channel, wherein the second rate is higher than the first rate, and wherein the second subchannel is sent after the first subchannel.

136. The method of claim 126, further comprising:

if signaling information for the user terminal is not obtained from the second subchannel, receiving signaling information sent at a third rate on a third subchannel of the forward control channel, wherein the third rate is higher than the second rate, and wherein the third subchannel is sent after the second subchannel.

137. The method of claim 126, further comprising:

terminating processing of the forward control channel upon encountering decoding failure for a subchannel of the forward control channel.

138. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

a receive data processor operative to

receive signaling information sent at a first rate on a first subchannel of a forward control channel, and

if signaling information for the apparatus is not obtained from the first subchannel, receiving signaling information sent at a second rate on a second subchannel of the forward control channel, wherein the second rate is higher than the first rate, and wherein the second subchannel is sent after the first subchannel; and

a controller operative to direct the processing for the first and second subchannels.

139. The apparatus of claim 138, wherein the receive data processor is further operative to, if signaling information for the apparatus is not obtained from the second subchannel, receive signaling information sent at a third rate on a third subchannel of the forward control channel, wherein the third rate is higher than the second rate, and wherein the third subchannel is sent after the second subchannel.

140. The apparatus of claim 138, wherein the controller is further operative to terminate processing of the forward control channel upon encountering decoding failure for a subchannel of the forward control channel.

141. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

means for receiving signaling information sent at a first rate on a first subchannel of a forward control channel; and

means for, if signaling information for the apparatus is not obtained from the first subchannel, receiving signaling information sent at a second rate on a second subchannel of the forward control channel, wherein the second rate is higher than the first rate, and wherein the second subchannel is sent after the first subchannel.

142. The apparatus of claim 141, further comprising:

means for, if signaling information for the apparatus is not obtained from the second subchannel, receiving signaling information sent at a third rate on a third subchannel of the forward control channel, wherein the third rate is higher than the second rate, and wherein the third subchannel is sent after the second subchannel.

143. The apparatus of claim 141, further comprising:

means for terminating processing of the forward control channel upon encountering decoding failure for a subchannel of the forward control channel.

144. A method of processing data for transmission in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

coding a data frame in accordance with a coding scheme to obtain a coded data frame;

partitioning the coded data frame into a plurality of coded data subframes, one coded data subframe for each of a plurality of spatial channels;

interleaving each coded data subframe in accordance with an interleaving scheme to obtain a corresponding interleaved data subframe, wherein a plurality of interleaved data subframes are obtained for the plurality of spatial channels; and

modulating each interleaved data subframe to obtain a corresponding stream of modulation symbols, wherein a plurality of modulation symbol streams are obtained for the plurality of spatial channels.

145. The method of claim 144, wherein the coded data frame is partitioned by completely filling one coded data subframe at a time.

146. The method of claim 144, wherein the coded data frame is partitioned by cycling through the plurality of coded data subframes for a plurality of iterations and partially filling each coded data subframe with a particular number of code bits from the coded data frame in each iteration.

147. The method of claim 144, wherein each of the plurality of spatial channels is associated with a respective rate, wherein the rate for each spatial channel indicates a particular modulation scheme and a particular code rate to use for the spatial channel, and wherein the particular modulation scheme is selected from among a plurality of modulation schemes supported by the system and the particular code rate is selected from among a plurality of code rates supported by the system.

148. The method of claim 147, wherein the plurality of code rates are obtained based on a single base code and a plurality of puncturing patterns.

149. The method of claim 144, further comprising:

puncturing each coded data subframe to obtain a code rate selected for the spatial channel for the coded data subframe.

150. The method of claim 144, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM).

151. The method of claim 150, wherein the coded data frame is partitioned by cycling through the plurality of coded data subframes for a plurality of iterations and partially filling each coded data subframe with code bits, from the coded data frame, for a group of M subbands in each iteration, where M is greater than one and less than a total number of subbands used for data transmission.

152. The method of claim 151, wherein the interleaving is performed for the code bits for each group of M subbands.

153. The method of claim 150, further comprising:

processing the plurality of modulation symbol streams to obtain a plurality of streams of OFDM symbols, wherein the OFDM symbols have a cyclic prefix length selected from among at least two cyclic prefix lengths supported by the system.

154. The method of claim 150, further comprising:

processing the plurality of modulation symbol streams to obtain a plurality of streams of OFDM symbols, wherein the OFDM symbols are of a size selected from among at least two OFDM symbol sizes supported by the system.

155. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

an encoder operative to code a data frame in accordance with a coding scheme to obtain a coded data frame;

a demultiplexer operative to partition the coded data frame into a plurality of coded data subframes, one coded data subframe for each of a plurality of spatial channels;

an interleaver operative to interleave each coded data subframe in accordance with an interleaving scheme to obtain a corresponding interleaved data subframe, wherein a plurality of interleaved data subframes are obtained for the plurality of spatial channels; and

a symbol mapping unit operative to modulate each interleaved data subframe to obtain a corresponding stream of modulation symbols, wherein a plurality of modulation symbol streams are obtained for the plurality of spatial channels.

156. The apparatus of claim 155, wherein each of the plurality of spatial channels is associated with a respective rate, wherein the rate for each spatial channel indicates a particular modulation scheme and a particular code rate to use for the spatial channel, and wherein the particular modulation scheme is selected from among a plurality of modulation schemes supported by the system and the particular code rate is selected from among a plurality of code rates supported by the system.

157. The apparatus of claim 155, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM).

158. The apparatus of claim 157, wherein the demultiplexer is operative to partition the coded data frame by cycling through the plurality of coded data subframes for a plurality of iterations and partially filling each coded data subframe with code bits,

from the coded data frame, for a group of M subbands in each iteration, where M is greater than one and less than a total number of subbands used for data transmission.

159. The apparatus of claim 157, further comprising:

a plurality of OFDM modulators operative to process the plurality of modulation symbol streams to obtain a plurality of streams of OFDM symbols, wherein the OFDM symbols have a cyclic prefix length selected from among at least two cyclic prefix lengths supported by the system.

160. The apparatus of claim 157, further comprising:

a plurality of OFDM modulators operative to process the plurality of modulation symbol streams to obtain a plurality of streams of OFDM symbols, wherein the OFDM symbols are of a size selected from among at least two OFDM symbol sizes supported by the system.

161. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

means for coding a data frame in accordance with a coding scheme to obtain a coded data frame;

means for partitioning the coded data frame into a plurality of coded data subframes, one coded data subframe for each of a plurality of spatial channels;

means for interleaving each coded data subframe in accordance with an interleaving scheme to obtain a corresponding interleaved data subframe, wherein a plurality of interleaved data subframes are obtained for the plurality of spatial channels; and

means for modulating each interleaved data subframe to obtain a corresponding stream of modulation symbols, wherein a plurality of modulation symbol streams are obtained for the plurality of spatial channels.

162. The apparatus of claim 161, wherein each of the plurality of spatial channels is associated with a respective rate, wherein the rate for each spatial channel indicates a particular modulation scheme and a particular code rate to use for the spatial channel, and wherein the particular modulation scheme is selected from among a

plurality of modulation schemes supported by the system and the particular code rate is selected from among a plurality of code rates supported by the system.

163. The apparatus of claim 161, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM).

164. The apparatus of claim 163, wherein the coded data frame is partitioned by cycling through the plurality of coded data subframes for a plurality of iterations and partially filling each coded data subframe with code bits, from the coded data frame, for a group of M subbands in each iteration, where M is greater than one and less than a total number of subbands used for data transmission.

165. The apparatus of claim 163, further comprising:

means for processing the plurality of modulation symbol streams to obtain a plurality of streams of OFDM symbols, wherein the OFDM symbols have a cyclic prefix length selected from among at least two cyclic prefix lengths supported by the system.

166. The apparatus of claim 163, further comprising:

means for processing the plurality of modulation symbol streams to obtain a plurality of streams of OFDM symbols, wherein the OFDM symbols are of a size selected from among at least two OFDM symbol sizes supported by the system.

167. A method of accessing a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

receiving system information via a first transport channel on a downlink;

transmitting an access request via a second transport channel on an uplink, wherein the access request is transmitted based on the received system information;

monitoring a third transport channel on the downlink for an acknowledgment of the transmitted access request; and

repeating the receiving, transmitting, and monitoring if the acknowledgment is not received within a predetermined time period.

168. The method of claim 167, wherein the monitoring the third transport channel includes

monitoring an acknowledgment bit in the first transport channel, and

processing the third transport channel for the acknowledgment if the acknowledgment bit is set.

169. The method of claim 167, wherein a plurality of access requests are transmitted.

170. The method of claim 169, wherein the plurality of access requests are transmitted at successively lower rates.

171. The method of claim 169, further comprising:

waiting a pseudo-random period of time prior to transmitting a next access request among the plurality of access requests.

172. The method of claim 169, further comprising:

transmitting a steered pilot along with the access request on the second transport channel, the steered pilot being sent on at least one eigenmode of a MIMO channel for the uplink.

173. The method of claim 167, wherein the system information indicates a time interval in which transmission of access requests is allowed, and wherein the access request is transmitted within the time interval.

174. The method of claim 167, wherein the system information indicates a particular number of slots in which transmission of access requests is allowed, and wherein the access request identifies a specific slot in which the access request is transmitted.

175. The method of claim 174, wherein the slots have a time duration configurable by the system.

176. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a receive data processor operative to receive system information via a first transport channel on a downlink;

a transmit data processor operative to process an access request for transmission via a second transport channel on an uplink, wherein the access request is transmitted based on the received system information;

a controller operative to monitor a third transport channel on the downlink for an acknowledgment of the transmitted access request, and

wherein the receive data processor is operative to receive updated system information, the transmit data processor operative to process another access request, and the controller operative to monitor the third transport channel if the acknowledgment is not received within a predetermined time period.

177. The apparatus of claim 176, wherein the controller is operative to monitor an acknowledgment bit in the first transport channel and to direct the receive data processor to process the third transport channel for the acknowledgment if the acknowledgment bit is set.

178. The apparatus of claim 176, wherein a plurality of access requests are transmitted.

179. The apparatus of claim 178, wherein the plurality of access requests are transmitted at successively lower rates.

180. The apparatus of claim 178, wherein the controller is operative to wait a pseudo-random period of time prior to initiating transmission of a next access request among the plurality of access requests.

181. The apparatus of claim 178, further comprising:

a transmit spatial processor operative to transmit a steered pilot along with the access request on the second transport channel, the steered pilot being sent on at least one eigenmode of a MIMO channel for the uplink.

182. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for receiving system information via a first transport channel on a downlink;

means for transmitting an access request via a second transport channel on an uplink, wherein the access request is transmitted based on the received system information;

means for monitoring a third transport channel on the downlink for an acknowledgment of the transmitted access request; and

means for repeating the receiving, transmitting, and monitoring if the acknowledgment is not received within a predetermined time period.

183. The apparatus of claim 182, wherein the means for monitoring the third transport channel includes

means for monitoring an acknowledgment bit in the first transport channel, and

means for processing the third transport channel for the acknowledgment if the acknowledgment bit is set.

184. The apparatus of claim 182, wherein a plurality of access requests are transmitted.

185. The apparatus of claim 184, wherein the plurality of access requests are transmitted at successively lower rates.

186. The apparatus of claim 184, further comprising:

means for waiting a pseudo-random period of time prior to transmitting a next access request among the plurality of access requests.

187. The apparatus of claim 184, further comprising:

means for transmitting a steered pilot along with the access request on the second transport channel, the steered pilot being sent on at least one eigenmode of a MIMO channel for the uplink.

188. A method of transmitting data in a wireless time division duplex (TDD) multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

estimating a channel response of a first communication link;

determining at least one rate for at least one spatial channel of a second communication link, one rate for each spatial channel, based on the estimated channel response of the first communication link; and

transmitting data on the at least one spatial channel of the second communication link at the at least one rate.

189. The method of claim 188, wherein the first communication link is an uplink and the second communication link is a downlink in the MIMO system.

190. The method of claim 188, further comprising:

estimating signal-to-noise-and-interference ratios (SNRs) of a plurality of spatial channels of the second communication link based on a noise estimate for the first communication link and the estimated channel response of the first communication link; and

selecting the at least one spatial channel from among the plurality of spatial channels based on the SNRs of the plurality of spatial channels.

191. The method of claim 190, wherein the at least one spatial channel is further selected based on a water-filling procedure, and wherein the at least one rate is determined based on SNR of the at least one spatial channel and the water-filling procedure.

192. The method of claim 188, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM)..

193. The method of claim 192, wherein a plurality of spatial channels are obtained for each of a plurality of subbands, wherein a plurality of wideband spatial channels are formed by the plurality of spatial channels of the plurality of subbands,

each wideband spatial channel including one spatial channel of each of the plurality of subbands.

194. The method of claim 193, wherein at least one wideband spatial channel is selected for data transmission based on SNRs for the plurality of spatial channels of the plurality of subbands.

195. The method of claim 194, wherein the at least one wideband spatial channel is further selected based on channel inversion to achieve similar SNRs across the plurality of subbands of each wideband spatial channel.

196. The method of claim 188, wherein the channel response of the first communication link is estimated based on a steered pilot received via a plurality of eigenmodes of the first communication link, and wherein each of the at least one spatial channel corresponds to one of the plurality of eigenmodes.

197. An apparatus in a wireless time division duplex (TDD) multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to estimate a channel response of a first communication link and to determine at least one rate for at least one spatial channel of a second communication link, one rate for each spatial channel, based on the estimated channel response of the first communication link; and

a transmit data processor operative to process data based on the at least one rate for transmission on the at least one spatial channel of the second communication link.

198. The apparatus of claim 197, wherein the controller is further operative to estimate signal-to-noise-and-interference ratios (SNRs) of a plurality of spatial channels of the second communication link based on a noise estimate for the first communication link and the estimated channel response of the first communication link, and to select the at least one spatial channel from among the plurality of spatial channels based on the SNRs of the plurality of spatial channels.

199. The apparatus of claim 198, wherein the controller is further operative to select the at least one spatial channel based on a water-filling procedure, and to determine the at least one rate based on SNR of the at least one spatial channel and the water-filling procedure.

200. The apparatus of claim 197, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM).

201. The apparatus of claim 200, wherein a plurality of spatial channels are obtained for each of a plurality of subbands, wherein a plurality of wideband spatial channels are formed by the plurality of spatial channels of the plurality of subbands, each wideband spatial channel including one spatial channel of each of the plurality of subbands.

202. An apparatus in a wireless time division duplex (TDD) multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for estimating a channel response of a first communication link;

means for determining at least one rate for at least one spatial channel of a second communication link, one rate for each spatial channel, based on the estimated channel response of the first communication link; and

means for transmitting data on the at least one spatial channel of the second communication link at the at least one rate.

203. The apparatus of claim 202, further comprising:

means for estimating signal-to-noise-and-interference ratios (SNRs) of a plurality of spatial channels of the second communication link based on a noise estimate for the first communication link and the estimated channel response of the first communication link; and

means for selecting the at least one spatial channel from among the plurality of spatial channels based on the SNRs of the plurality of spatial channels.

204. The apparatus of claim 203, wherein the at least one spatial channel is further selected based on a water-filling procedure, and wherein the at least one rate is

determined based on SNR of the at least one spatial channel and the water-filling procedure.

205. The apparatus of claim 202, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM).

206. The apparatus of claim 205, wherein a plurality of spatial channels are obtained for each of a plurality of subbands, wherein a plurality of wideband spatial channels are formed by the plurality of spatial channels of the plurality of subbands, each wideband spatial channel including one spatial channel of each of the plurality of subbands.

207. A method of transmitting data in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

estimating a channel response of a first communication link;

determining at least one supported rate for at least one spatial channel of the first communication link based on the estimated channel response, one supported rate for each spatial channel, each supported rate indicating a maximum rate supported by the corresponding spatial channel for a predetermined level of performance;

sending the at least one supported rate via a second communication link to a transmitting entity;

receiving at least one selected rate for the at least one spatial channel, one selected rate for each spatial channel, each selected rate being equal to or less than the supported rate for the spatial channel; and

receiving data transmission on the at least one spatial channel of the first communication link at the at least one selected rate.

208. The method of claim 207, wherein the first communication link is an uplink and the second communication link is a downlink in the MIMO system.

209. The method of claim 207, wherein the estimated channel response of the first communication link includes signal-to-noise-and-interference ratios (SNRs) for a plurality of spatial channels of the first communication link, and wherein the at least one

spatial channel is selected from among the plurality of spatial channels based on the SNRs for the plurality of spatial channels.

210. The method of claim 209, wherein the at least one spatial channel is further selected based on a water-filling procedure, and wherein the at least one supported rate is determined based on SNR for the at least one spatial channel and the water-filling procedure.

211. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to estimate a channel response of a first communication link and to determine at least one supported rate for at least one spatial channel of the first communication link based on the estimated channel response, one supported rate for each spatial channel, each supported rate indicating a maximum rate supported by the corresponding spatial channel for a predetermined level of performance;

a transmit data processor operative to send the at least one supported rate via a second communication link to a transmitting entity;

a receive data processor operative to

receive at least one selected rate for the at least one spatial channel, one selected rate for each spatial channel, each selected rate being equal to or less than the supported rate for the spatial channel, and

process data transmission received on the at least one spatial channel of the first communication link at the at least one selected rate.

212. The apparatus of claim 211, wherein the estimated channel response of the first communication link includes signal-to-noise-and-interference ratios (SNRs) for a plurality of spatial channels of the first communication link, and wherein the controller is operative to select the at least one spatial channel from among the plurality of spatial channels based on the SNRs for the plurality of spatial channels.

213. The apparatus of claim 212, wherein the controller is operative to further select the at least one spatial channel based on a water-filling procedure and to

determine the at least one supported rate based on SNR for the at least one spatial channel and the water-filling procedure.

214. An apparatus in a wireless multiple-access multiple-input multiple-output (MIMO) communication system, comprising:

means for estimating channel response of a first communication link;

means for determining at least one supported rate for at least one spatial channel of the first communication link based on the estimated channel response, one supported rate for each spatial channel, each supported rate indicating a maximum rate supported by the corresponding spatial channel for a predetermined level of performance;

means for sending the at least one supported rate via a second communication link to a transmitting entity;

means for receiving at least one selected rate for the at least one spatial channel, one selected rate for each spatial channel, each selected rate being equal to or less than the supported rate for the spatial channel; and

means for receiving data transmission on the at least one spatial channel of the first communication link at the at least one selected rate.

215. The apparatus of claim 214, wherein the estimated channel response of the first communication link includes signal-to-noise-and-interference ratios (SNRs) for a plurality of spatial channels of the first communication link, and wherein the at least one spatial channel is selected from among the plurality of spatial channels based on the SNRs for the plurality of spatial channels.

216. The apparatus of claim 215, wherein the at least one spatial channel is further selected based on a water-filling procedure, and wherein the at least one supported rate is determined based on SNR for the at least one spatial channel and the water-filling procedure.

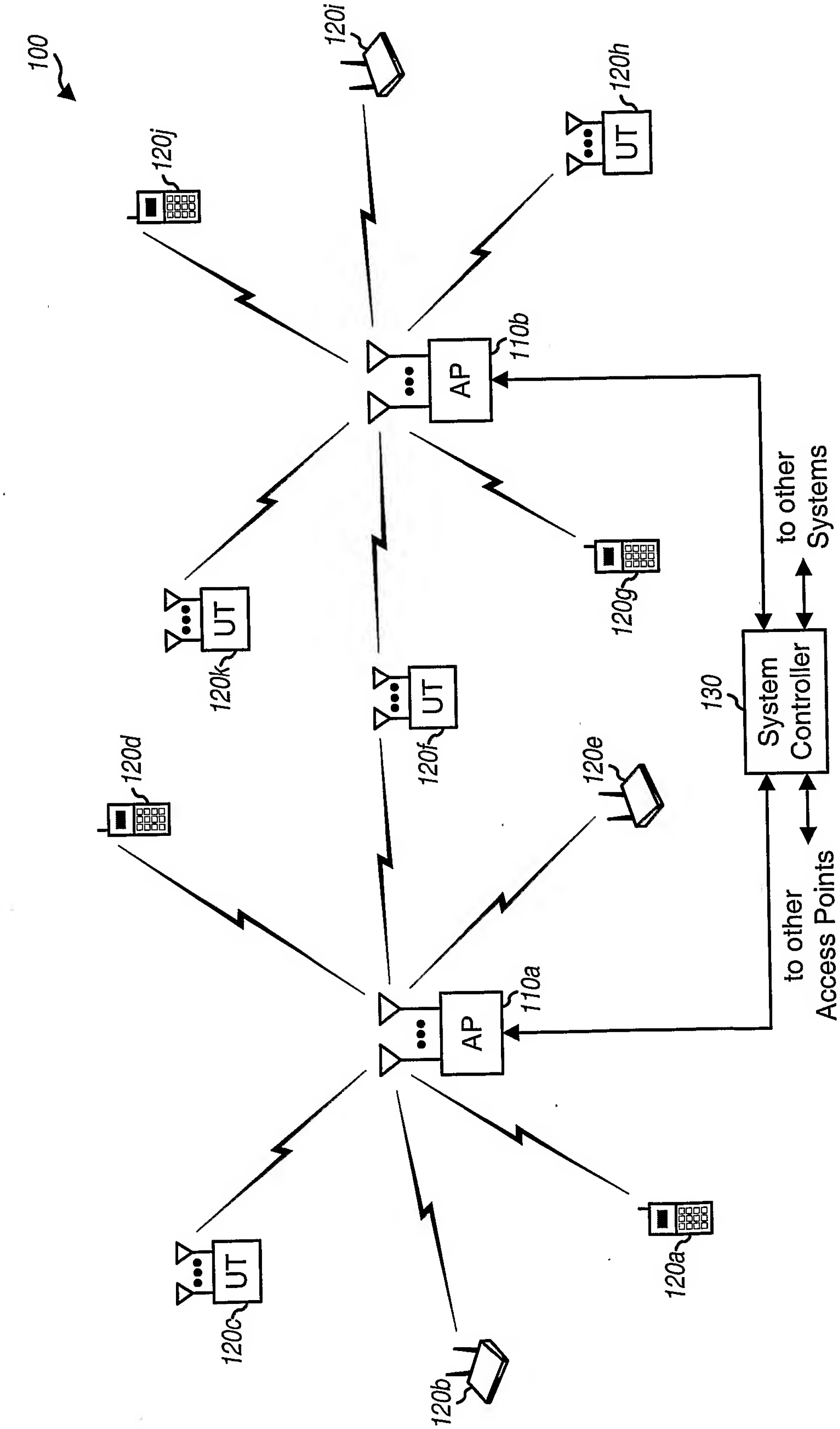

100
120i
120j
120h
UT
AP
110b
to other Systems
UT
120k
120g
UT
120f
120d
120e
System Controller
130
to other Access Points
AP
110a
UT
120c
120b
120a

FIG. 1

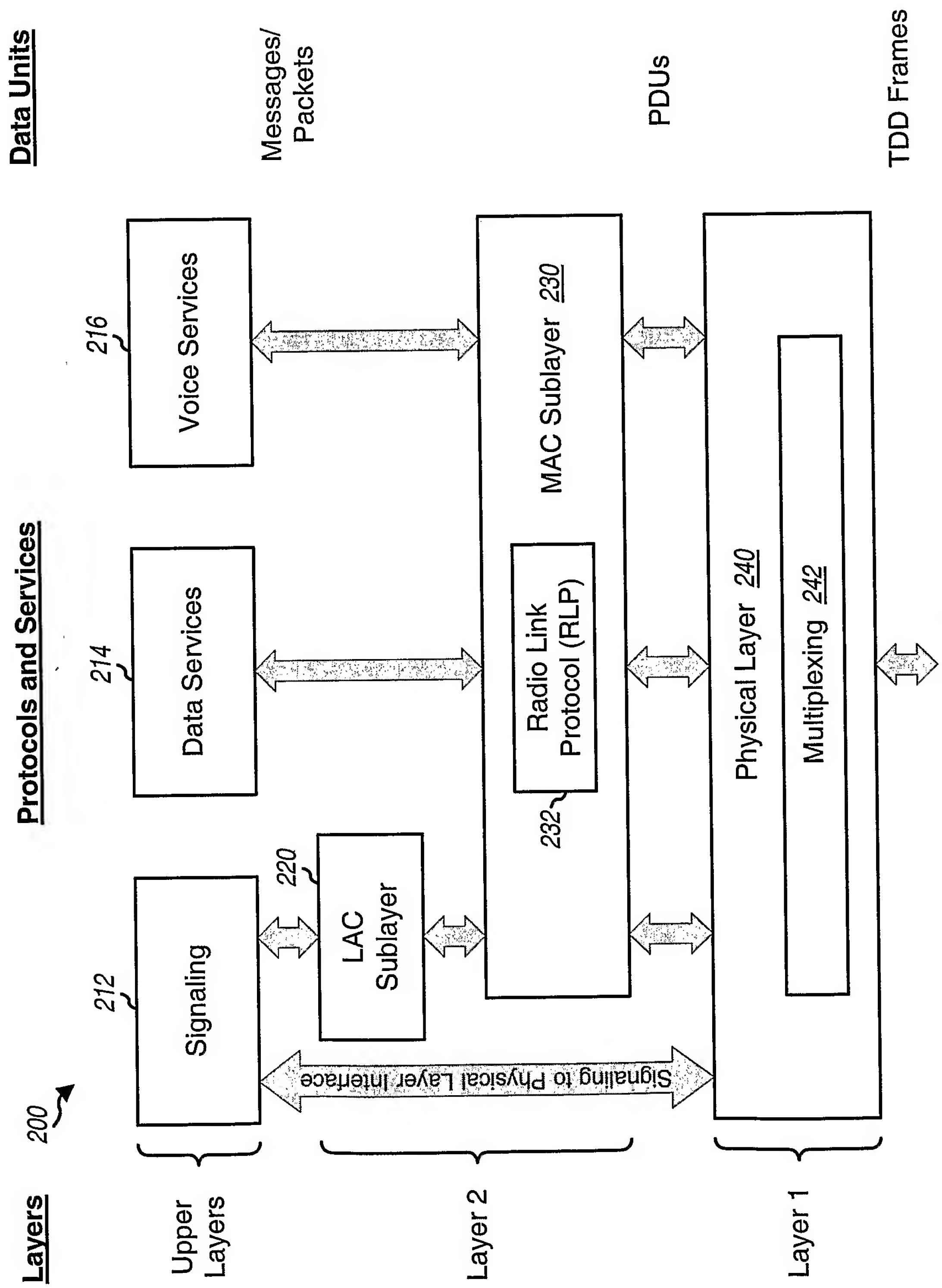

Layers
Protocols and Services
Data Units
200
Upper Layers
Layer 2
Layer 1
212
Signaling
214
Data Services
216
Voice Services
Messages/ Packets
220
LAC Sublayer
Signaling to Physical Layer Interface
MAC Sublayer 230
Radio Link Protocol (RLP)
232
PDUs
Physical Layer 240
Multiplexing 242
TDD Frames

FIG. 2

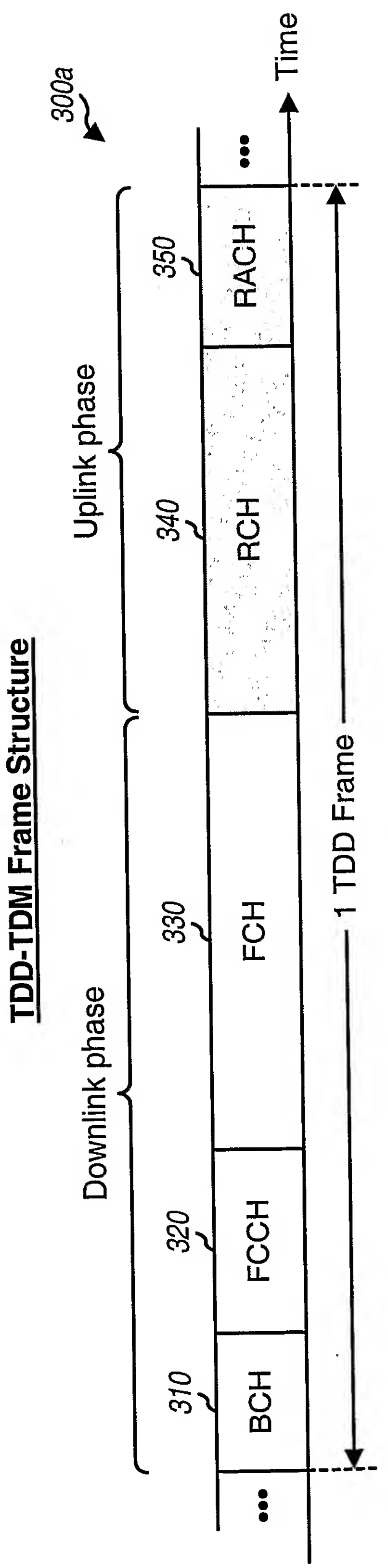

TDD-TDM Frame Structure
300a
Downlink phase
Uplink phase
310
BCH
320
FCCH
330
FCH
340
RCH
350
RACH
Time
1 TDD Frame

FIG. 3A

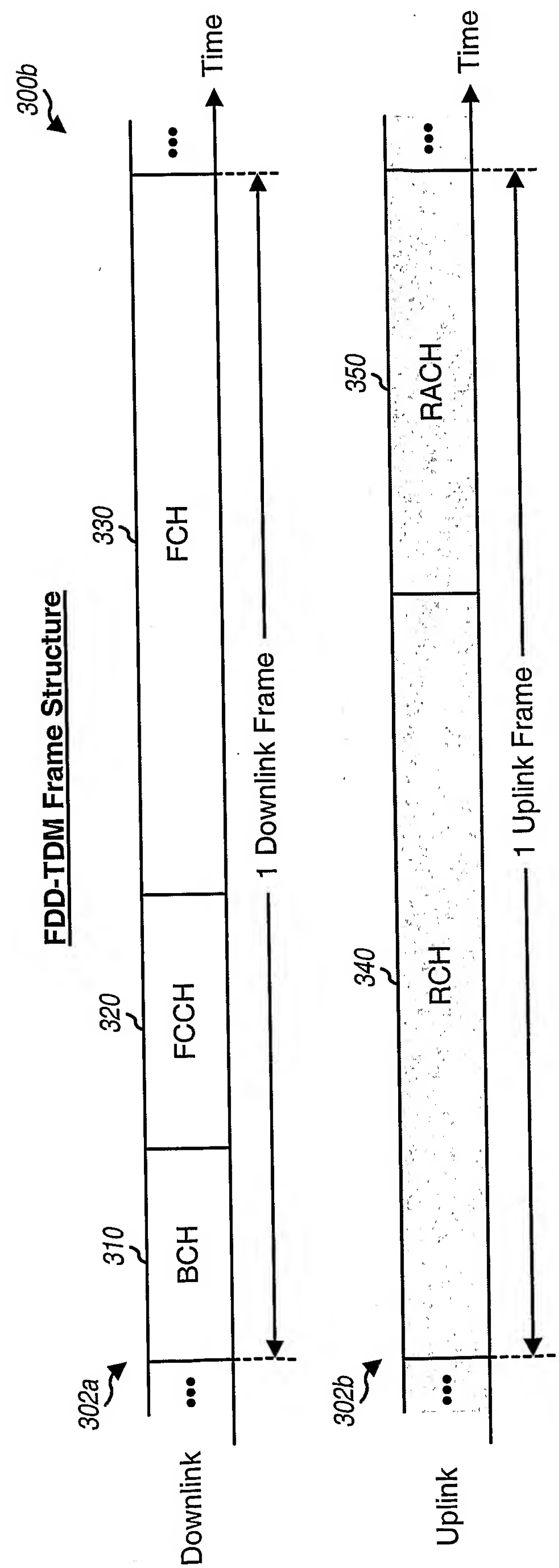

FDD-TDM Frame Structure
300b
Downlink
302a
310
BCH
320
FCCH
330
FCH
Time
1 Downlink Frame
Uplink
302b
340
RCH
350
RACH
Time
1 Uplink Frame

FIG. 3B

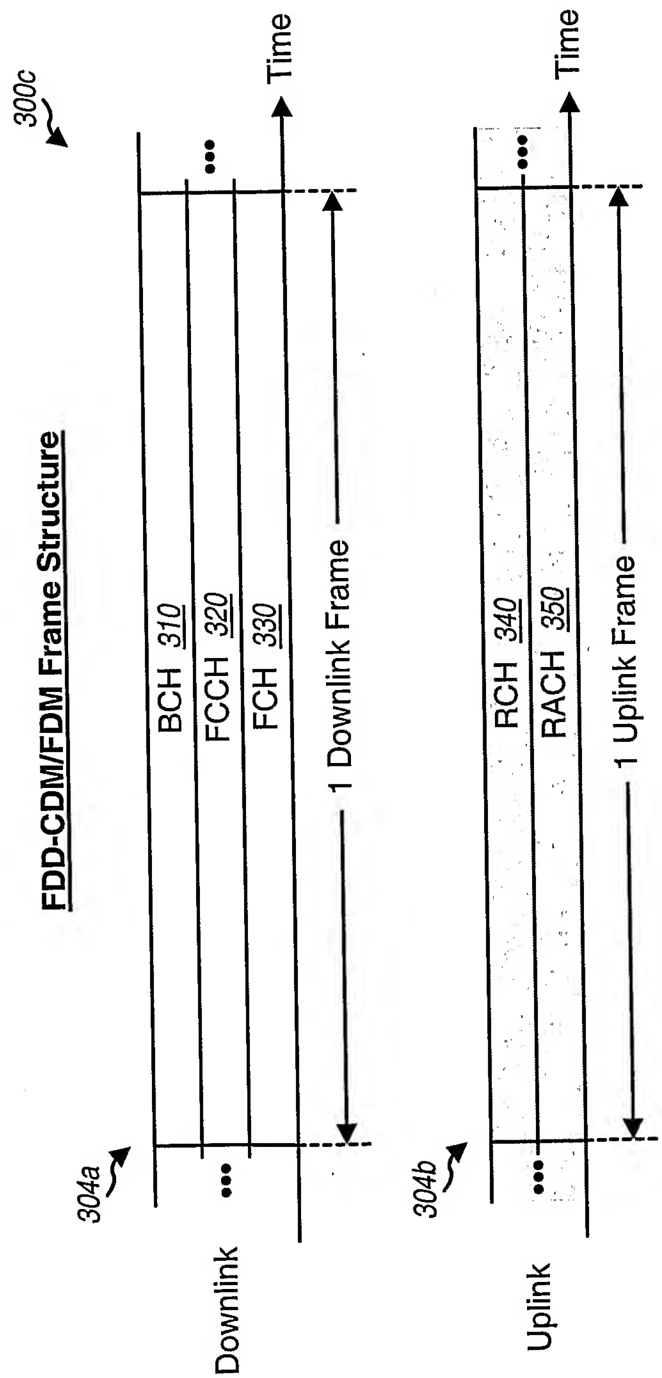

FDD-CDM/FDM Frame Structure
300c
Time
Time
BCH 310
FCCH 320
FCH 330
1 Downlink Frame
RCH 340
RACH 350
1 Uplink Frame
304a
304b
Downlink
Uplink

FIG. 3C

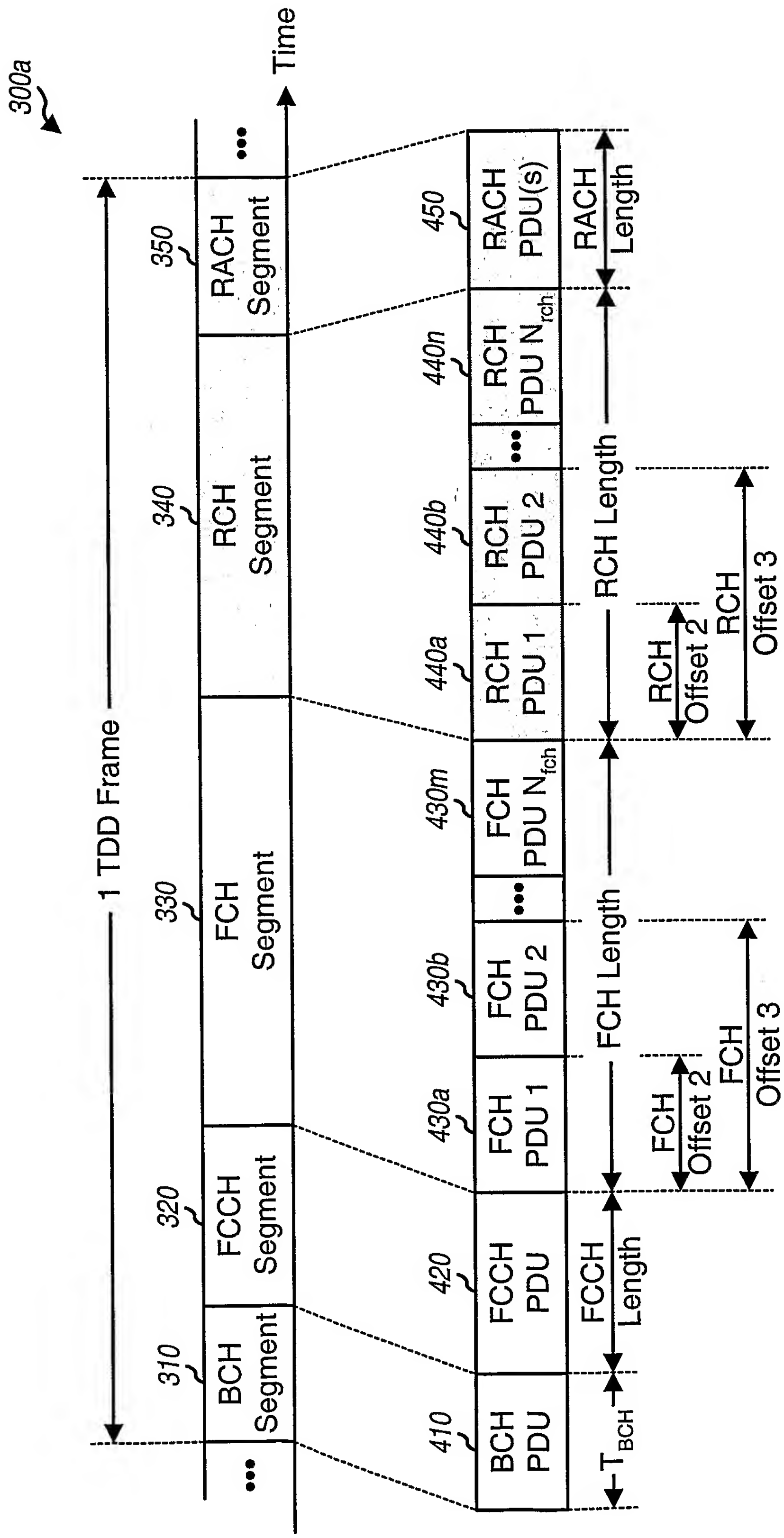

300a
Time
1 TDD Frame
310
BCH Segment
320
FCCH Segment
330
FCH Segment
340
RCH Segment
350
RACH Segment
410
BCH PDU
420
FCCH PDU
430a
FCH PDU 1
430b
FCH PDU 2
430m
FCH PDU N_fch
440a
RCH PDU 1
440b
RCH PDU 2
440n
RCH PDU N_rch
450
RACH PDU(s)
T_BCH
FCCH Length
FCH Length
RCH Length
RACH Length
FCH Offset 2
FCH Offset 3
RCH Offset 2
RCH Offset 3

FIG. 4

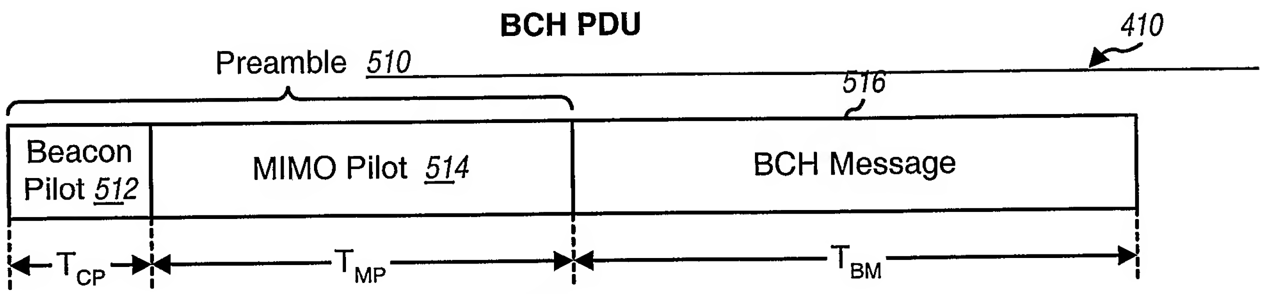


FIG. 5A

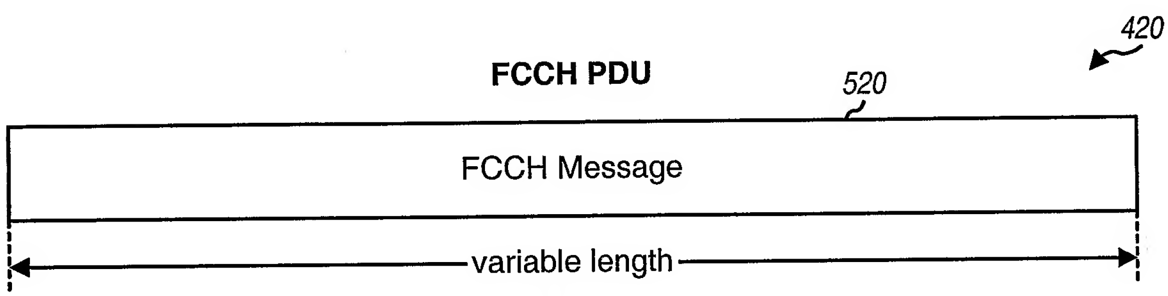


FIG. 5B

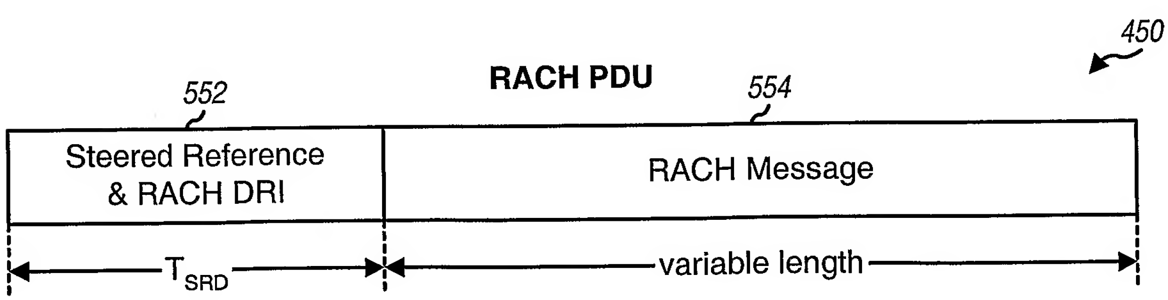


FIG. 5C

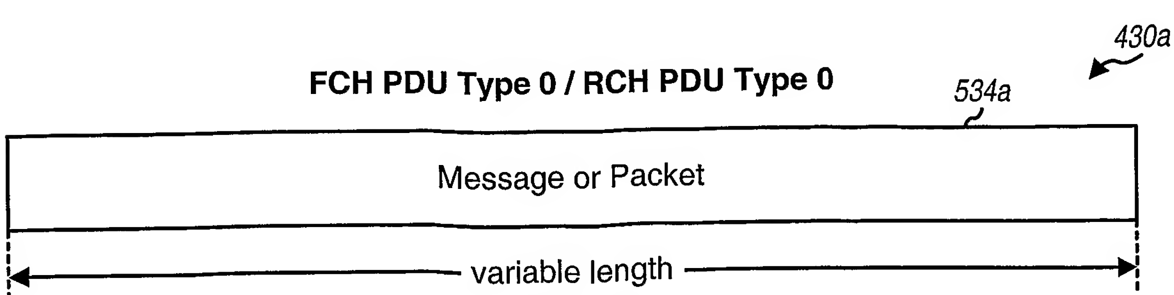


FIG. 5D

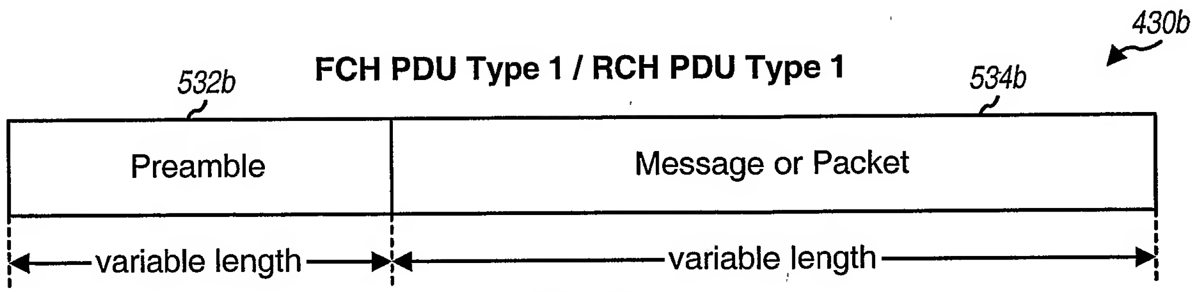


FIG. 5E

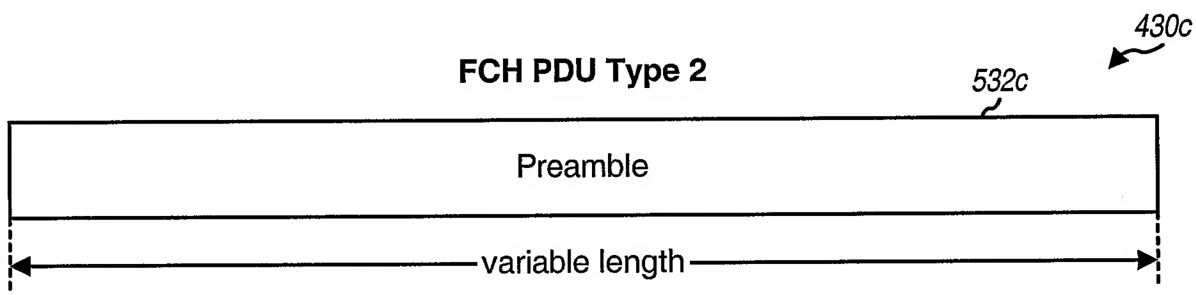


FIG. 5F

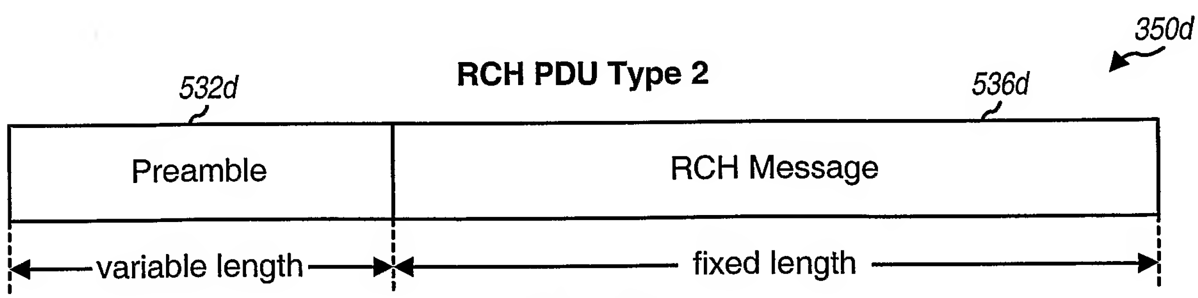


FIG. 5G

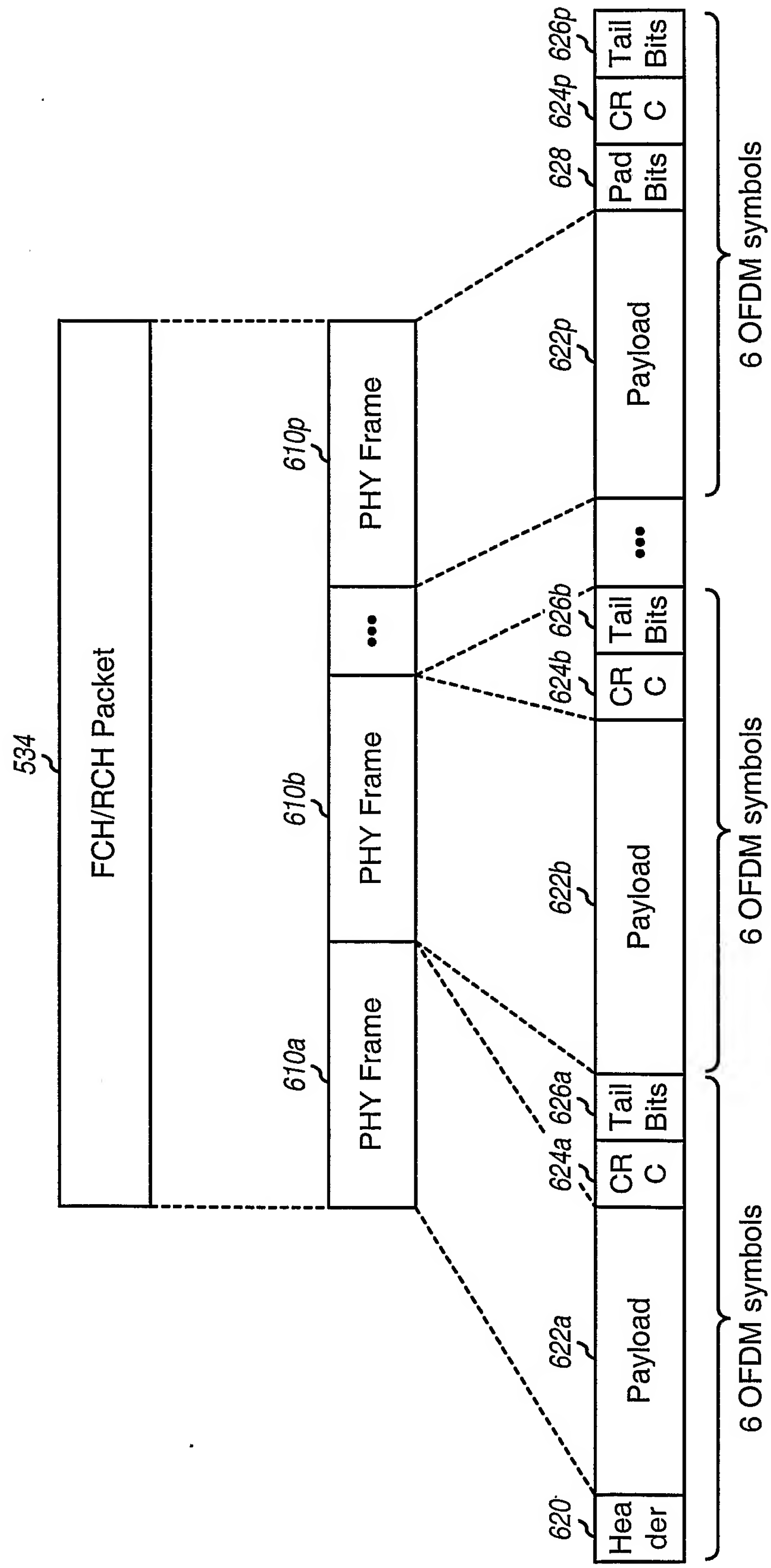

534
FCH/RCH Packet
610a
PHY Frame
610b
PHY Frame
610p
PHY Frame
620
Hea der
622a
Payload
624a
CR C
626a
Tail Bits
622b
Payload
624b
CR C
626b
Tail Bits
622p
Payload
628
Pad Bits
624p
CR C
626p
Tail Bits
6 OFDM symbols
6 OFDM symbols
6 OFDM symbols

FIG. 6

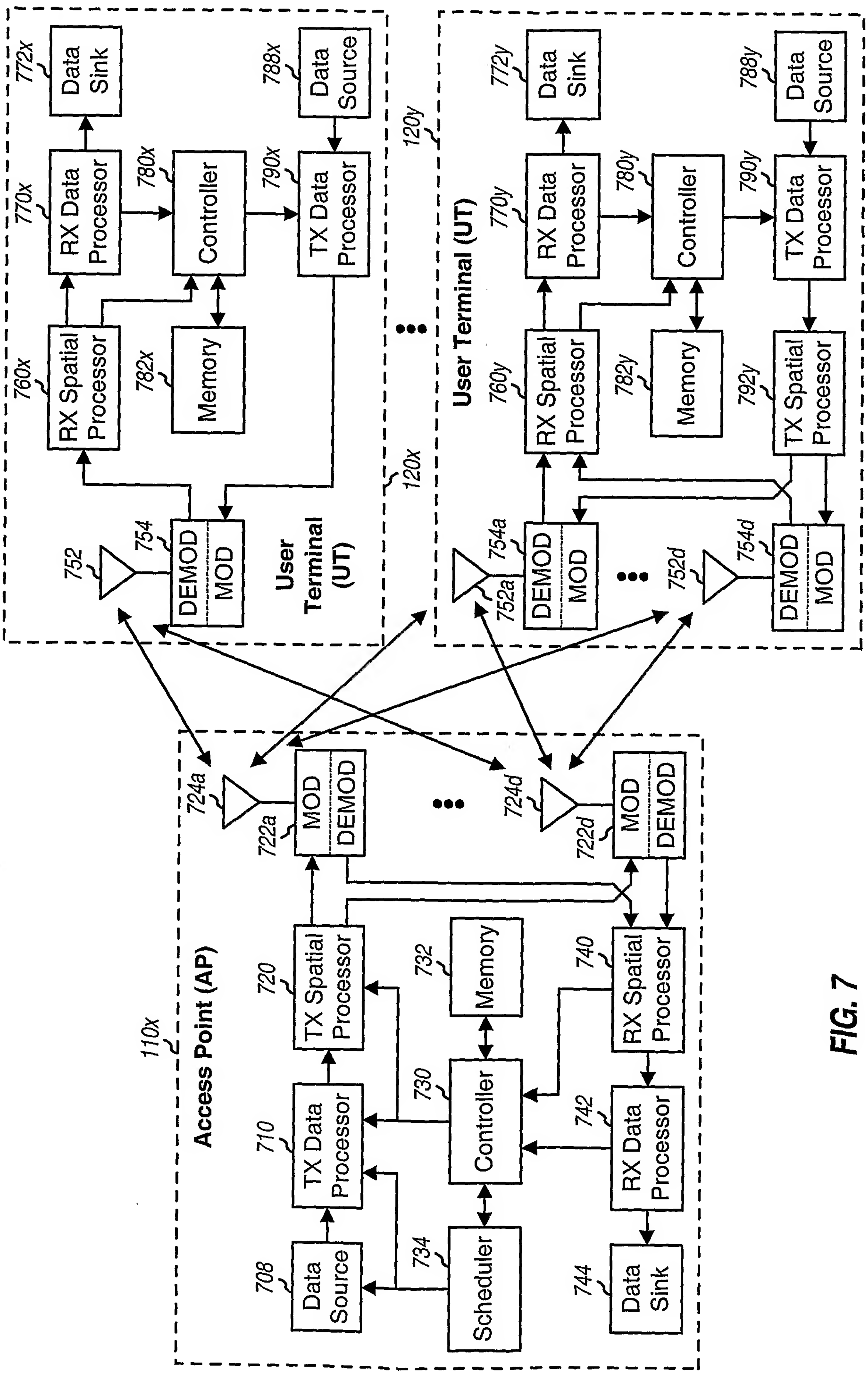

User Terminal (UT)
120x
752
754
DEMOD
MOD
760x
RX Spatial Processor
770x
RX Data Processor
772x
Data Sink
780x
Controller
782x
Memory
788x
Data Source
790x
TX Data Processor
User Terminal (UT)
120y
752a
754a
752d
754d
760y
RX Spatial Processor
770y
RX Data Processor
772y
Data Sink
780y
Controller
782y
Memory
788y
Data Source
790y
TX Data Processor
792y
TX Spatial Processor
Access Point (AP)
110x
708
Data Source
710
TX Data Processor
720
TX Spatial Processor
722a
724a
722d
724d
MOD
DEMOD
730
Controller
732
Memory
734
Scheduler
740
RX Spatial Processor
742
RX Data Processor
744
Data Sink

FIG. 7

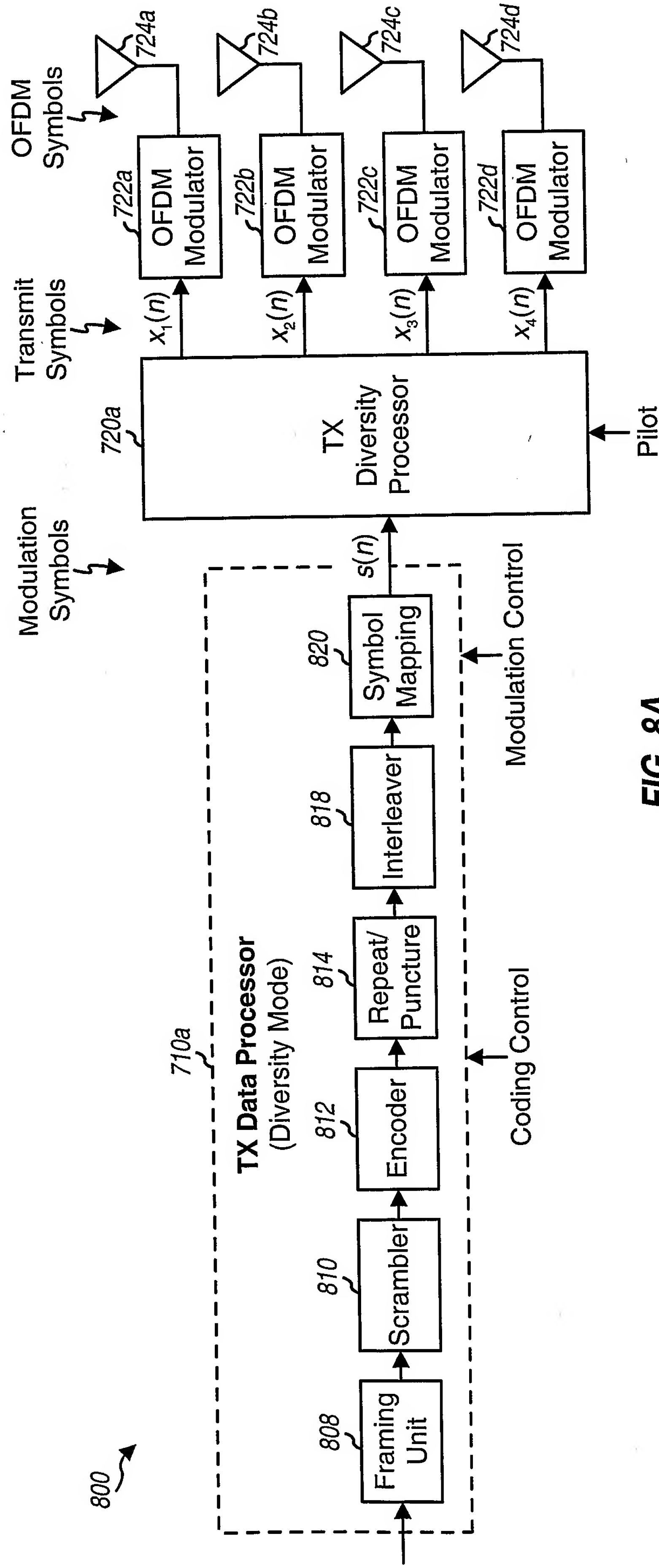


FIG. 8A

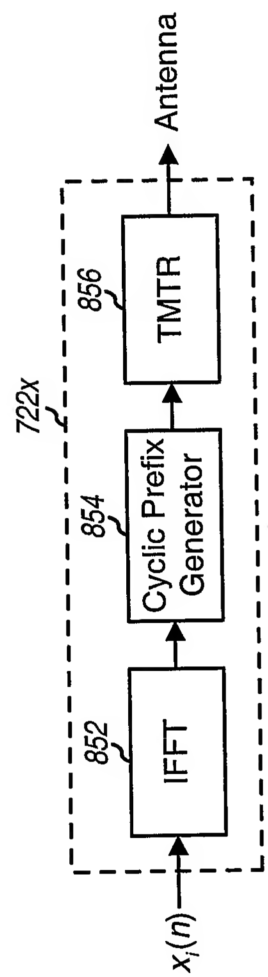


FIG. 8C

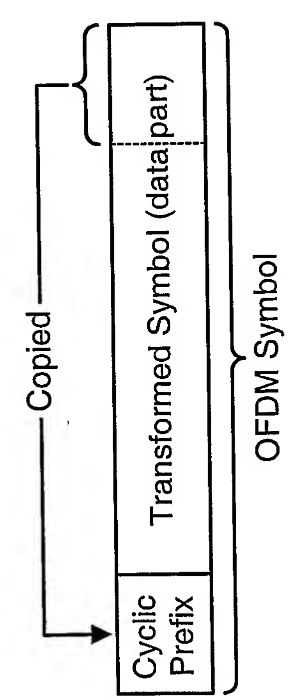


FIG. 8D

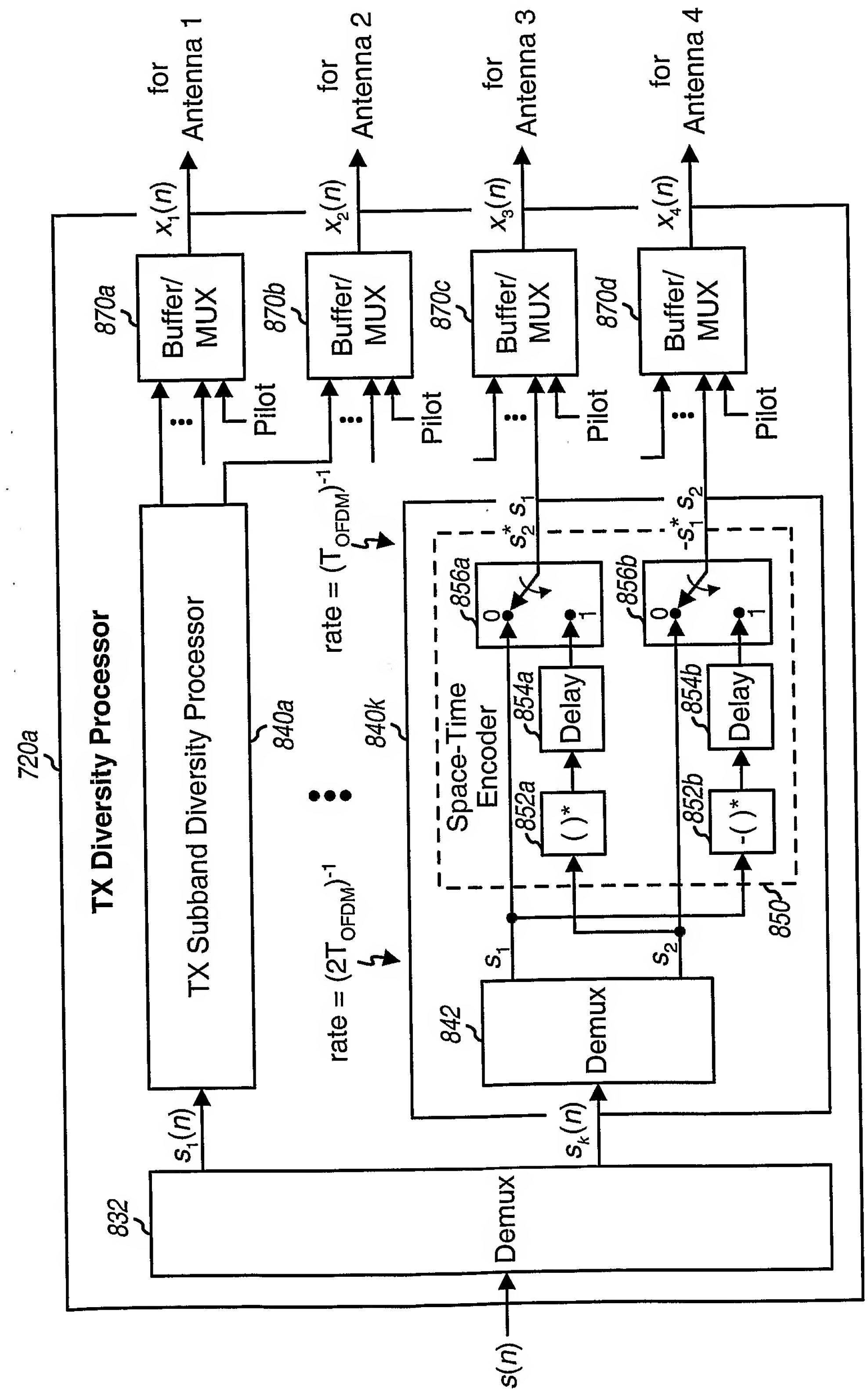
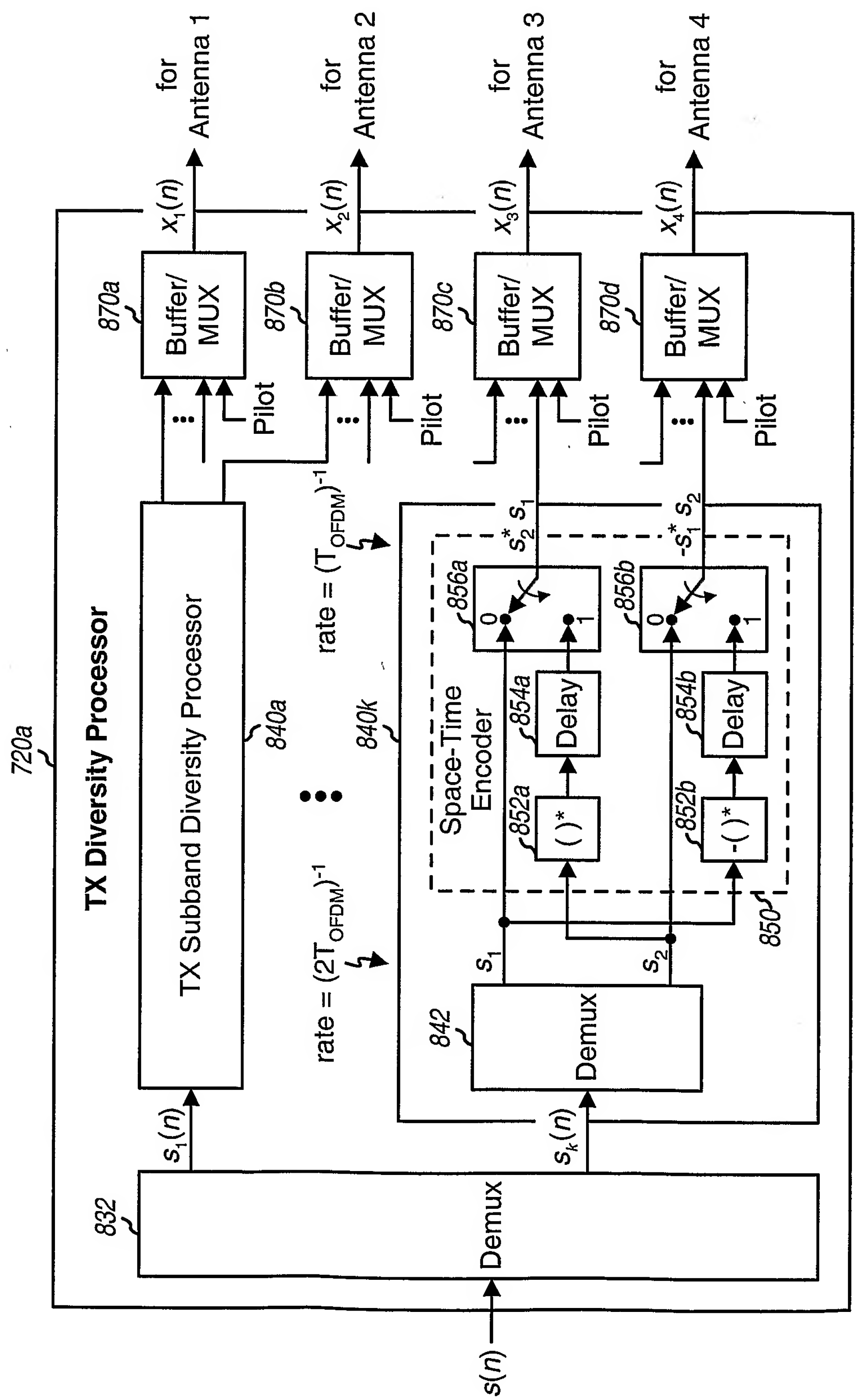

TX Diversity Processor
720a
832
Demux
s(n)
$s_1(n)$
$s_k(n)$
TX Subband Diversity Processor
840a
840k
rate = $(2T_{OFDM})^{-1}$
rate = $(T_{OFDM})^{-1}$
842
Demux
s_1
s_2
Space-Time Encoder
850
852a
( )*
852b
-( )*
854a
Delay
854b
Delay
856a
856b
0
1
$s_2^* \; s_1$
$-s_1^* \; s_2$
870a
870b
870c
870d
Buffer/ MUX
Pilot
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
for Antenna 1
for Antenna 2
for Antenna 3
for Antenna 4

FIG. 8B



720a
TX Diversity Processor
832
Demux
s(n)
$s_1(n)$
$s_k(n)$
TX Subband Diversity Processor
840a
840k
842
Demux
s_1
s_2
Space-Time Encoder
850
852a
( )*
852b
-( )*
854a
Delay
854b
Delay
856a
856b
0
1
$s_2^* \; s_1$
$-s_1^* \; s_2$
rate = $(2T_{OFDM})^{-1}$
rate = $(T_{OFDM})^{-1}$
Pilot
870a
870b
870c
870d
Buffer/ MUX
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
for Antenna 1
for Antenna 2
for Antenna 3
for Antenna 4

FIG. 8B

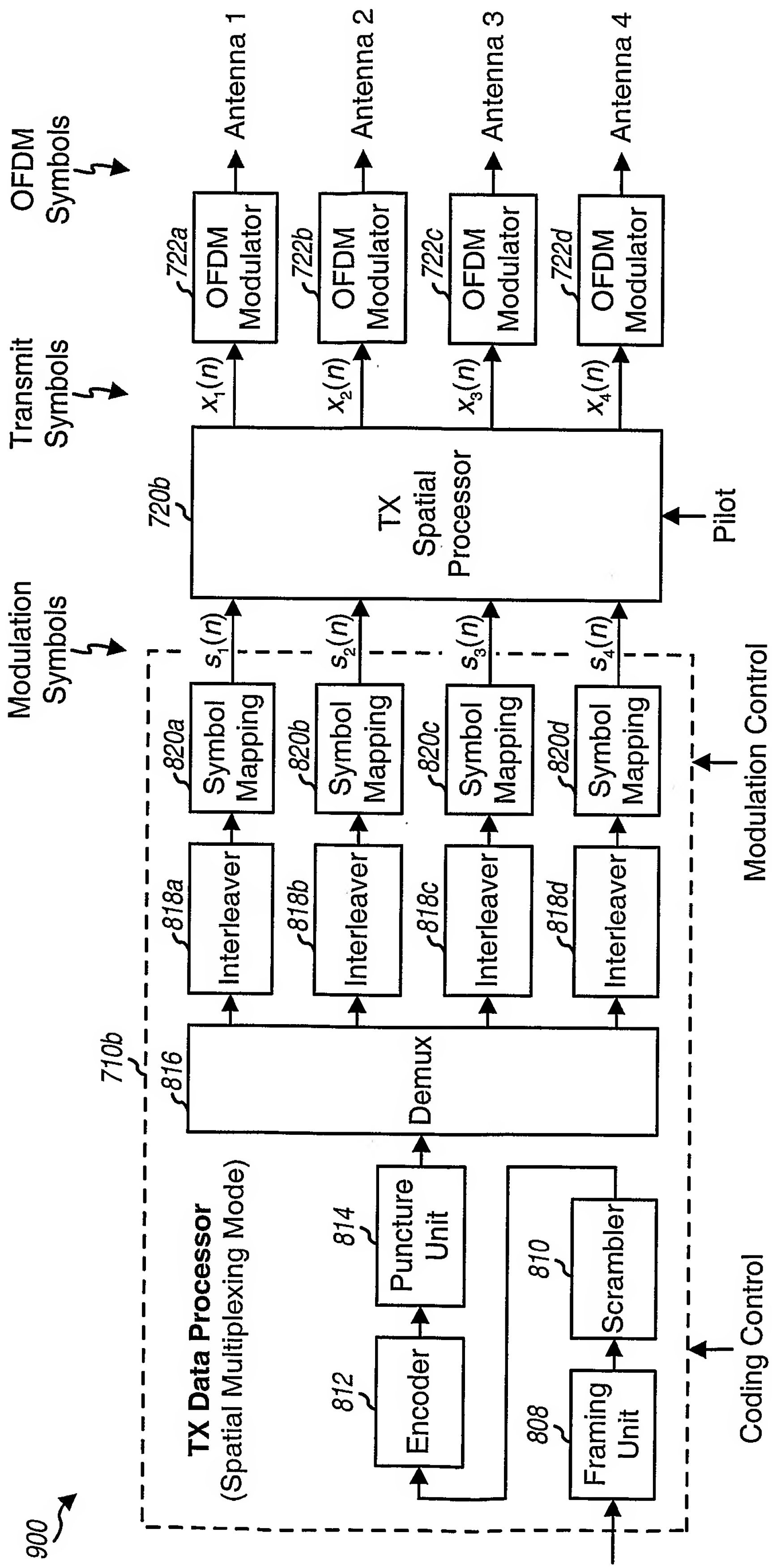
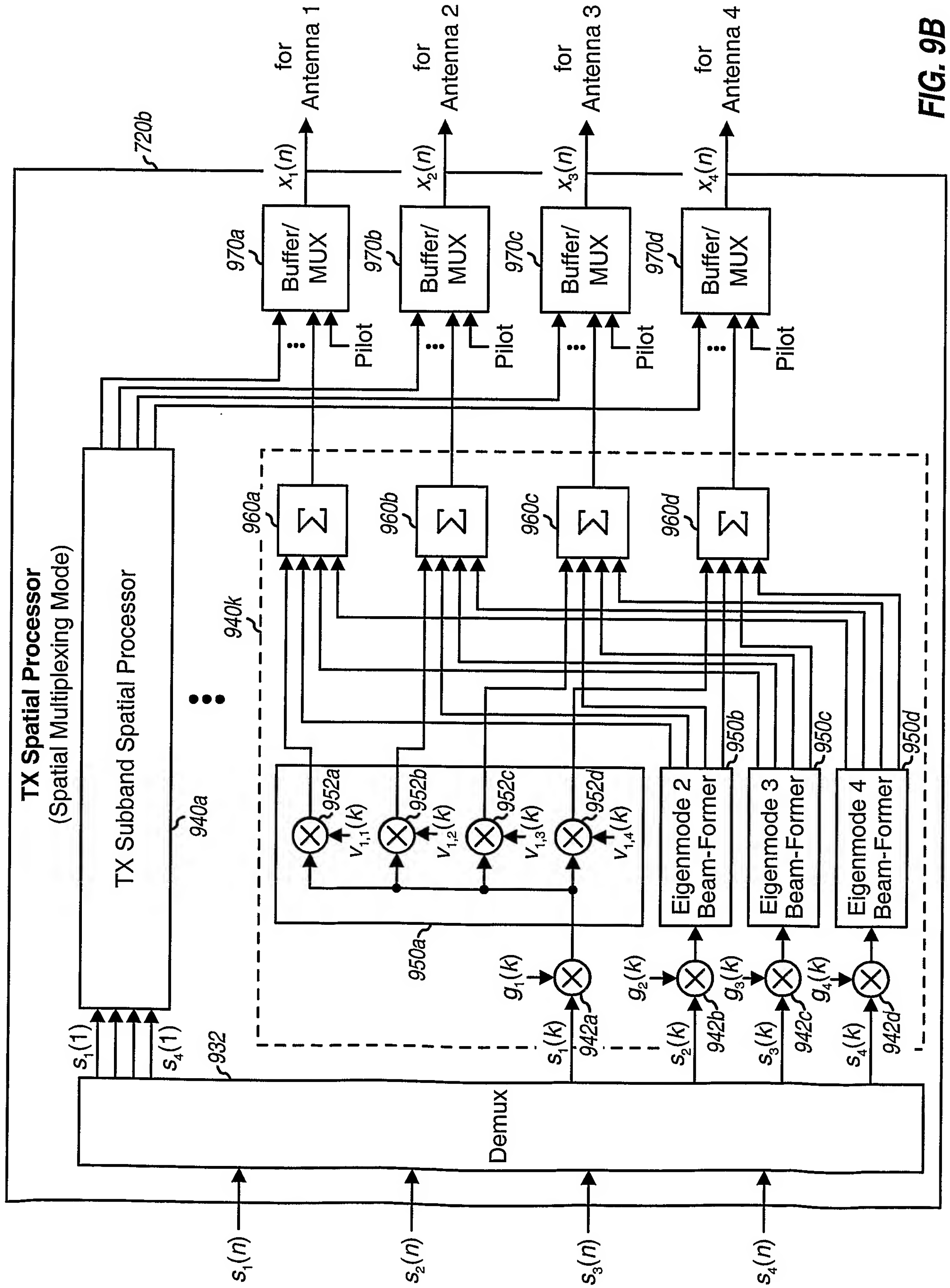

900
TX Data Processor
(Spatial Multiplexing Mode)
710b
808
Framing Unit
810
Scrambler
812
Encoder
814
Puncture Unit
816
Demux
818a
818b
818c
818d
Interleaver
Interleaver
Interleaver
Interleaver
820a
820b
820c
820d
Symbol Mapping
Symbol Mapping
Symbol Mapping
Symbol Mapping
Coding Control
Modulation Control
Modulation Symbols
$s_1(n)$
$s_2(n)$
$s_3(n)$
$s_4(n)$
720b
TX Spatial Processor
Pilot
Transmit Symbols
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
722a
722b
722c
722d
OFDM Modulator
OFDM Modulator
OFDM Modulator
OFDM Modulator
OFDM Symbols
Antenna 1
Antenna 2
Antenna 3
Antenna 4

FIG. 9A



TX Spatial Processor
(Spatial Multiplexing Mode)
720b
TX Subband Spatial Processor
940a
940k
Demux
932
$s_1(n)$
$s_2(n)$
$s_3(n)$
$s_4(n)$
$s_1(1)$
$s_4(1)$
$s_1(k)$
$s_2(k)$
$s_3(k)$
$s_4(k)$
942a
942b
942c
942d
$g_1(k)$
$g_2(k)$
$g_3(k)$
$g_4(k)$
950a
952a
952b
952c
952d
$v_{1,1}(k)$
$v_{1,2}(k)$
$v_{1,3}(k)$
$v_{1,4}(k)$
Eigenmode 2 Beam-Former
950b
Eigenmode 3 Beam-Former
950c
Eigenmode 4 Beam-Former
950d
960a
960b
960c
960d
Σ
Pilot
970a
970b
970c
970d
Buffer/ MUX
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
for Antenna 1
for Antenna 2
for Antenna 3
for Antenna 4

FIG. 9B

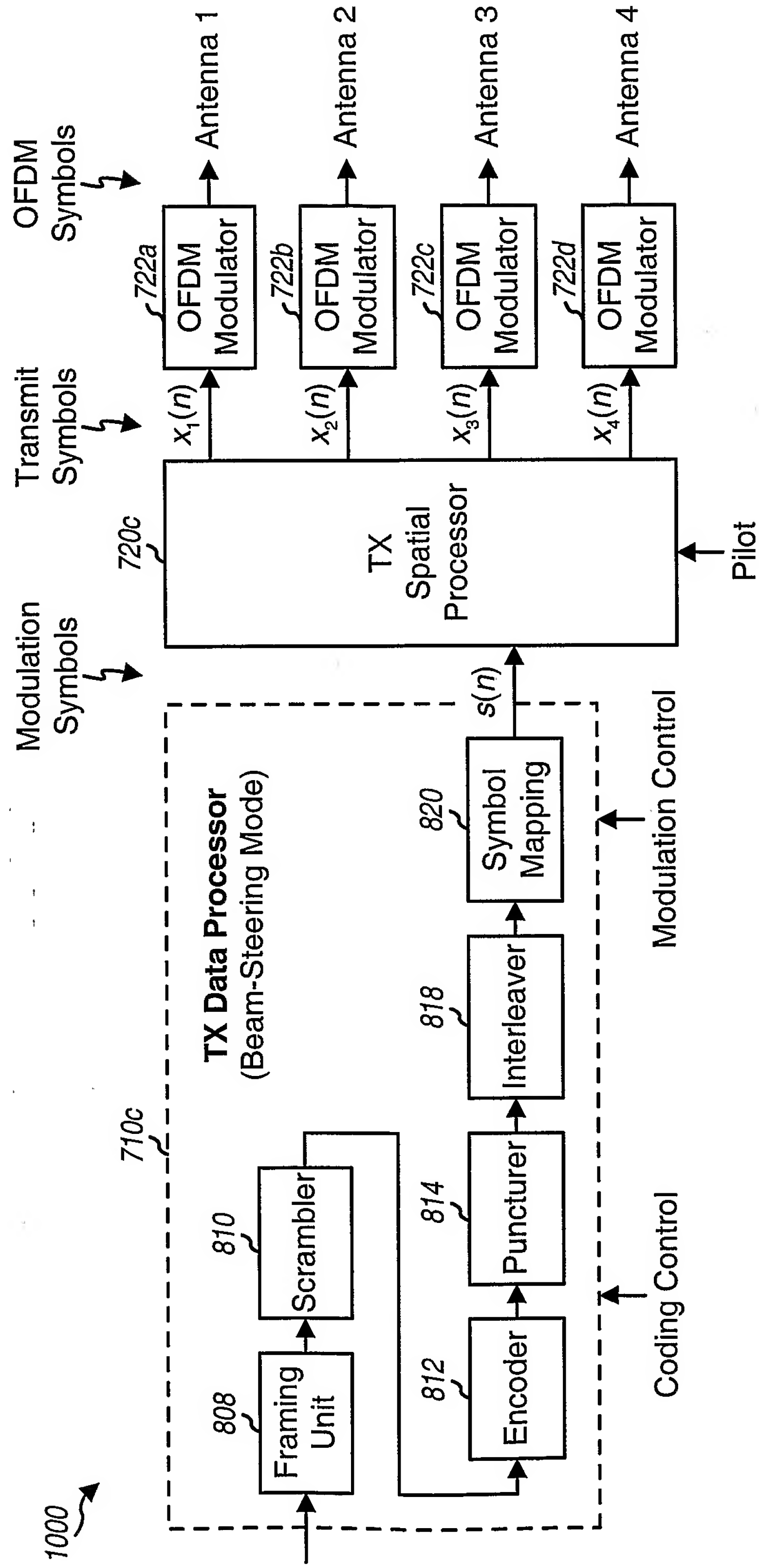

1000
TX Data Processor
(Beam-Steering Mode)
710c
808
Framing Unit
810
Scrambler
812
Encoder
814
Puncturer
818
Interleaver
820
Symbol Mapping
Coding Control
Modulation Control
s(n)
Modulation Symbols
720c
TX Spatial Processor
Pilot
Transmit Symbols
x1(n)
x2(n)
x3(n)
x4(n)
722a
OFDM Modulator
722b
OFDM Modulator
722c
OFDM Modulator
722d
OFDM Modulator
OFDM Symbols
Antenna 1
Antenna 2
Antenna 3
Antenna 4

FIG. 10A

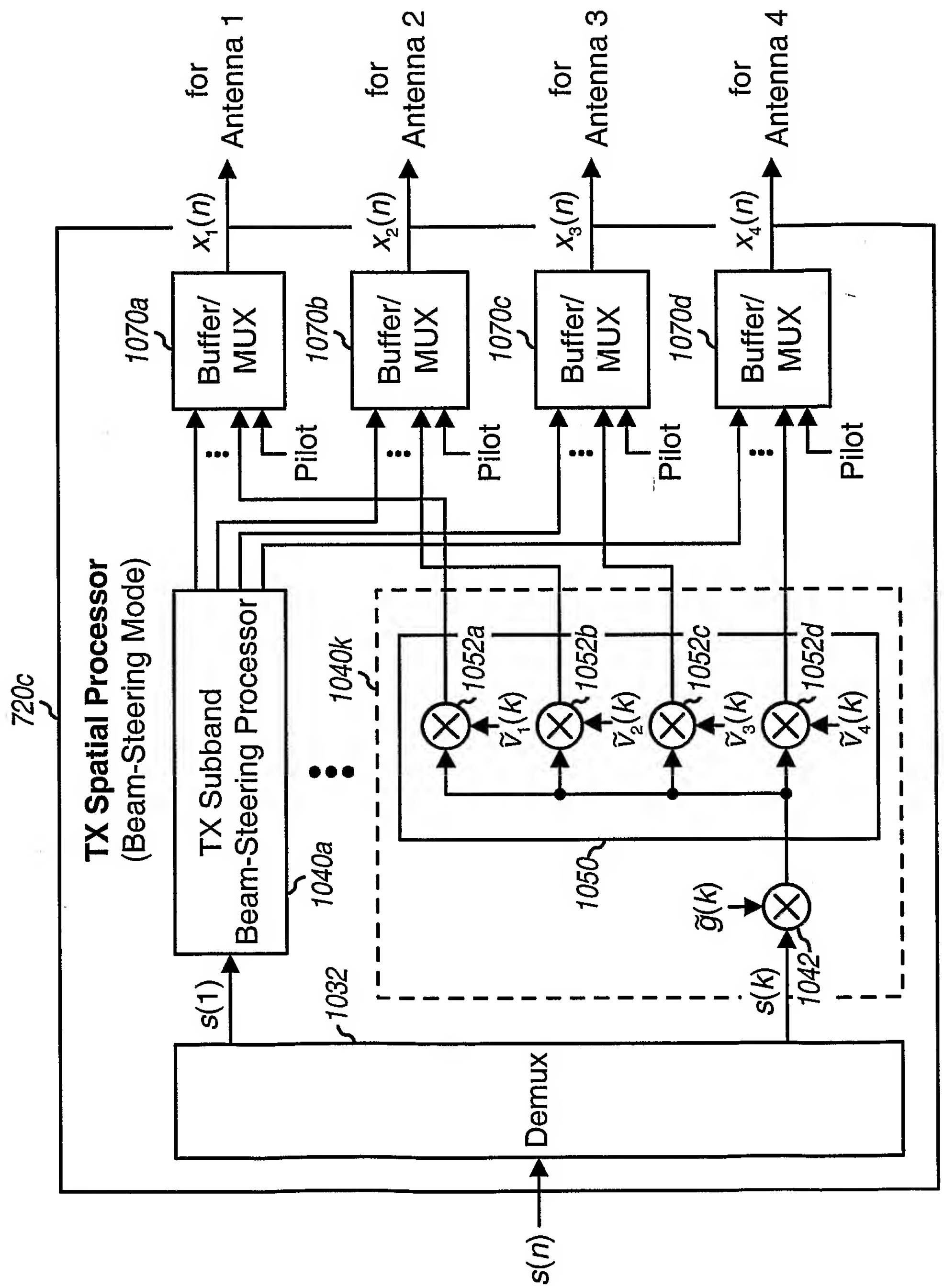

720c
TX Spatial Processor
(Beam-Steering Mode)
s(n)
Demux
1032
s(1)
TX Subband
Beam-Steering Processor
1040a
1040k
s(k)
1042
$\tilde{g}(k)$
1050
1052a
1052b
1052c
1052d
$\tilde{v}_1(k)$
$\tilde{v}_2(k)$
$\tilde{v}_3(k)$
$\tilde{v}_4(k)$
Pilot
Pilot
Pilot
Pilot
1070a
1070b
1070c
1070d
Buffer/
MUX
Buffer/
MUX
Buffer/
MUX
Buffer/
MUX
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
for
Antenna 1
for
Antenna 2
for
Antenna 3
for
Antenna 4

FIG. 10B

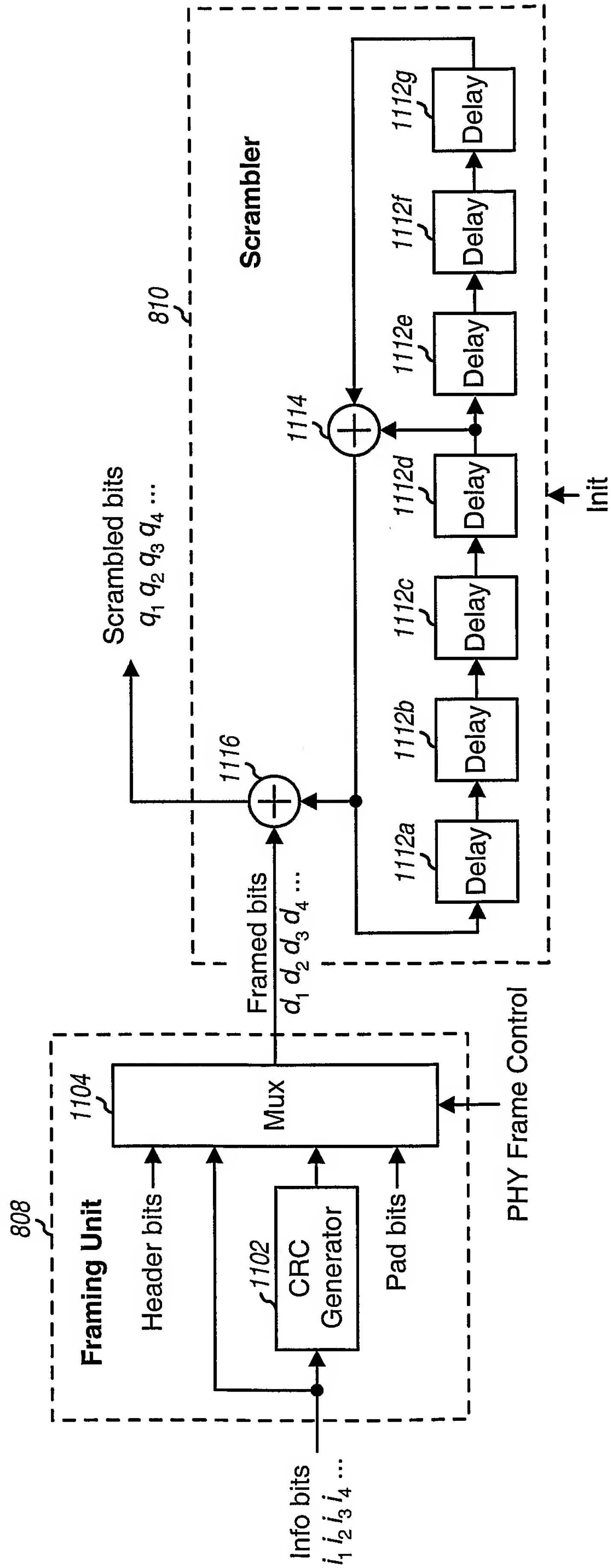

Info bits
$i_1\ i_2\ i_3\ i_4\ \ldots$
808
Framing Unit
Header bits
1102
CRC Generator
Pad bits
1104
Mux
PHY Frame Control
Framed bits
$d_1\ d_2\ d_3\ d_4\ \ldots$
810
Scrambler
1116
Scrambled bits
$q_1\ q_2\ q_3\ q_4\ \ldots$
1114
1112a
Delay
1112b
Delay
1112c
Delay
1112d
Delay
1112e
Delay
1112f
Delay
1112g
Delay
Init

FIG. 11A

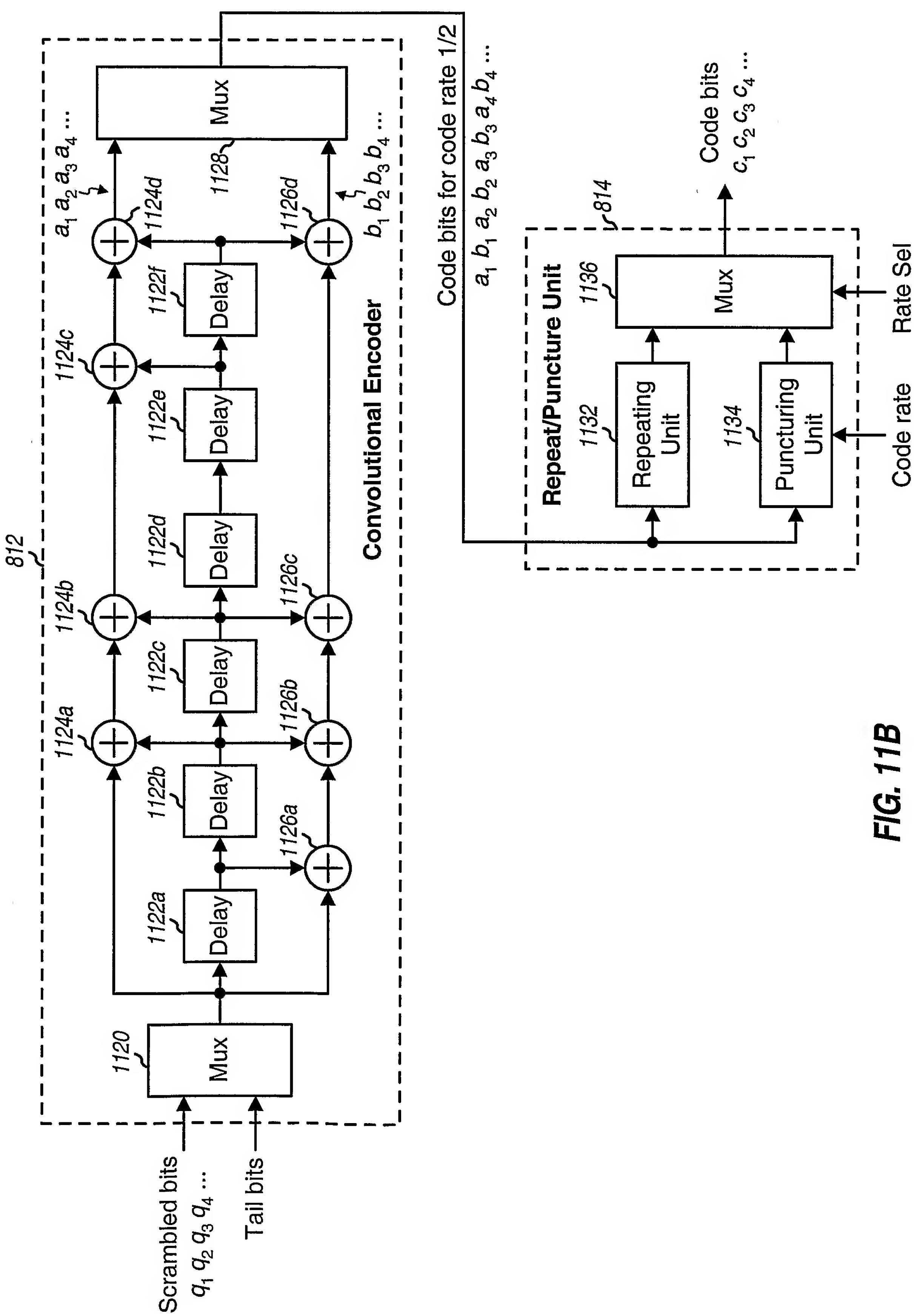

Scrambled bits
$q_1\ q_2\ q_3\ q_4$ ...
Tail bits
1120
Mux
Delay
1122a
1122b
1122c
1122d
1122e
1122f
1124a
1124b
1124c
1124d
1126a
1126b
1126c
1126d
812
Convolutional Encoder
1128
Mux
$a_1\ a_2\ a_3\ a_4$ ...
$b_1\ b_2\ b_3\ b_4$ ...
Code bits for code rate 1/2
$a_1\ b_1\ a_2\ b_2\ a_3\ b_3\ a_4\ b_4$ ...
814
Repeat/Puncture Unit
1132
Repeating Unit
1134
Puncturing Unit
1136
Mux
Code bits
$c_1\ c_2\ c_3\ c_4$ ...
Code rate
Rate Sel

FIG. 11B

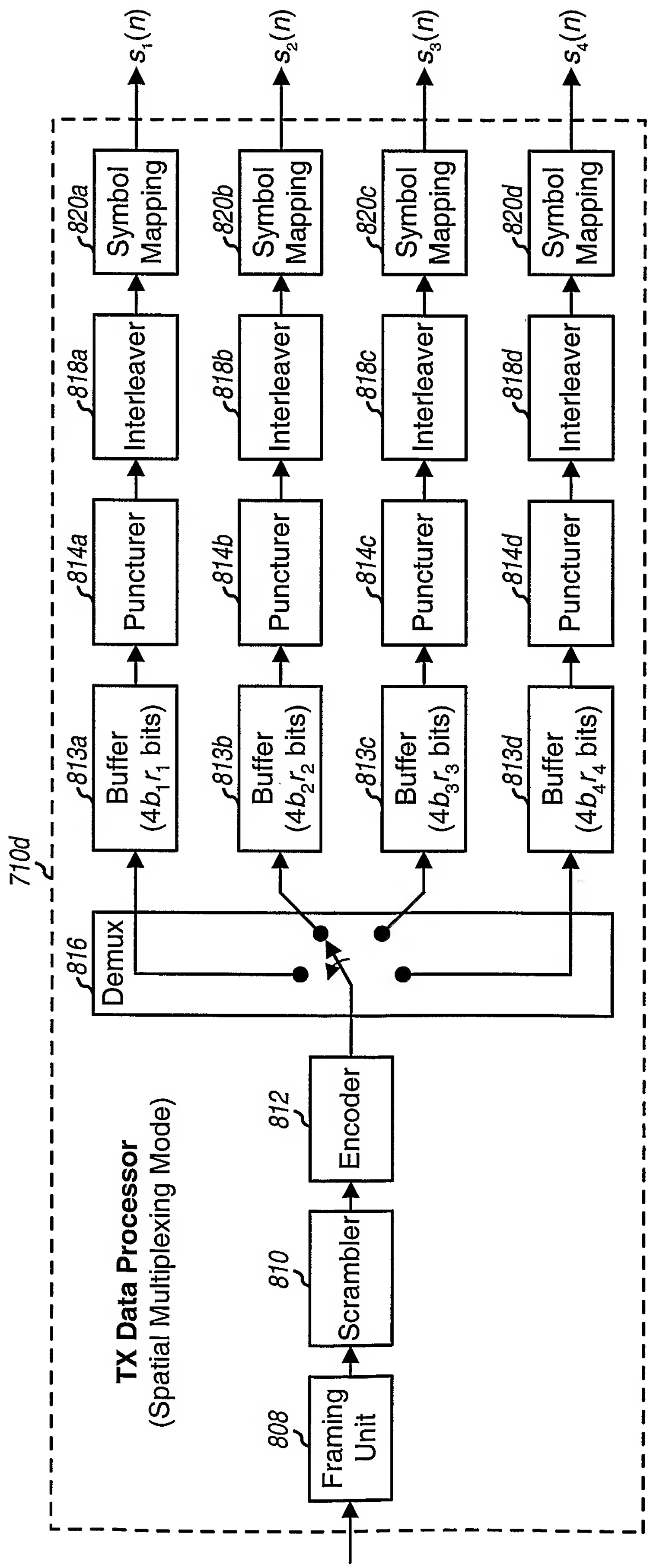

TX Data Processor
(Spatial Multiplexing Mode)
710d
808
Framing Unit
810
Scrambler
812
Encoder
816
Demux
813a
Buffer ($4b_1r_1$ bits)
813b
Buffer ($4b_2r_2$ bits)
813c
Buffer ($4b_3r_3$ bits)
813d
Buffer ($4b_4r_4$ bits)
814a
Puncturer
814b
Puncturer
814c
Puncturer
814d
Puncturer
818a
Interleaver
818b
Interleaver
818c
Interleaver
818d
Interleaver
820a
Symbol Mapping
820b
Symbol Mapping
820c
Symbol Mapping
820d
Symbol Mapping
$s_1(n)$
$s_2(n)$
$s_3(n)$
$s_4(n)$

FIG. 11C

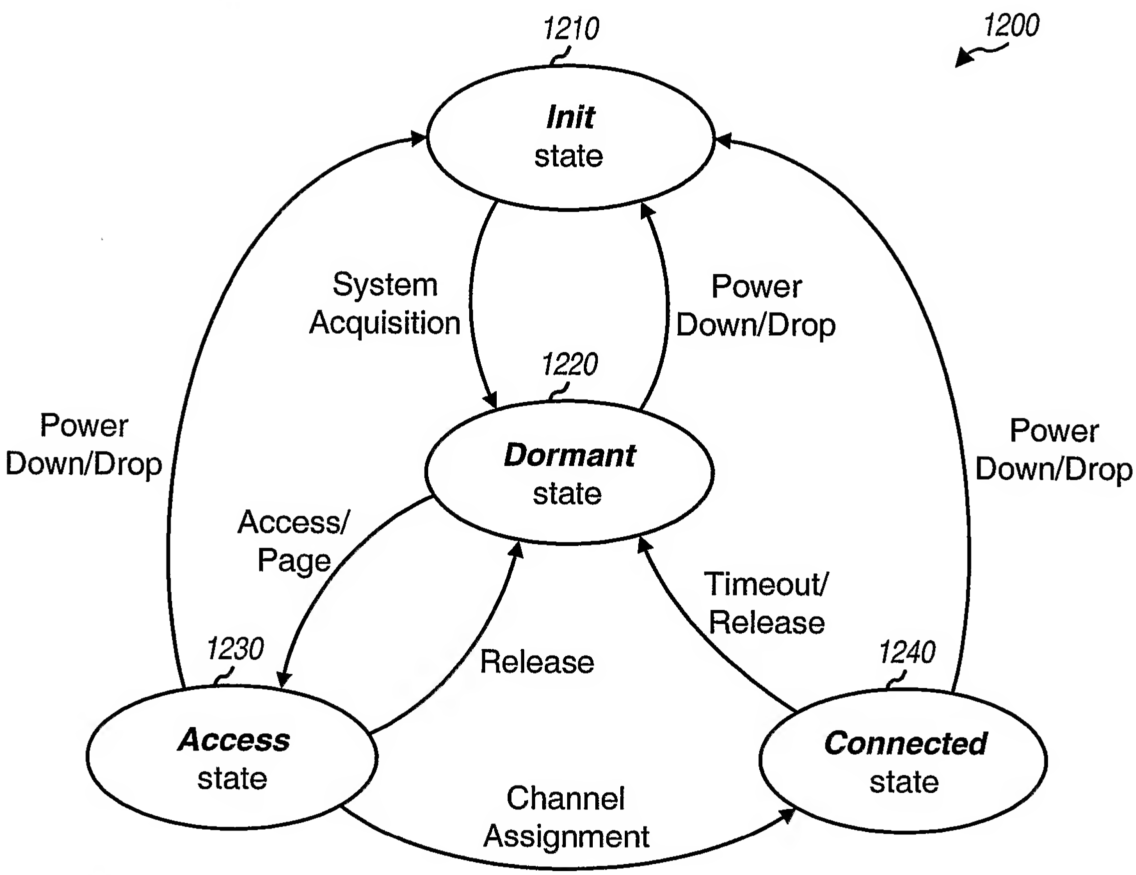


FIG. 12A

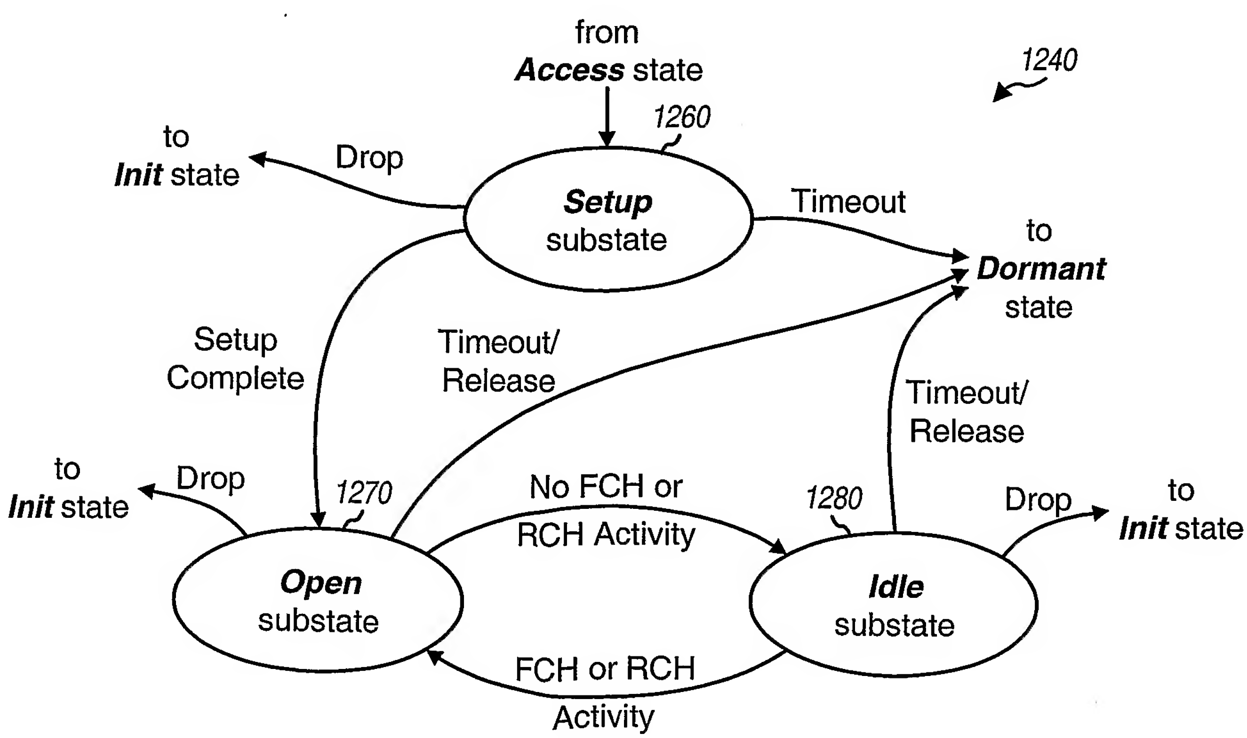


FIG. 12B

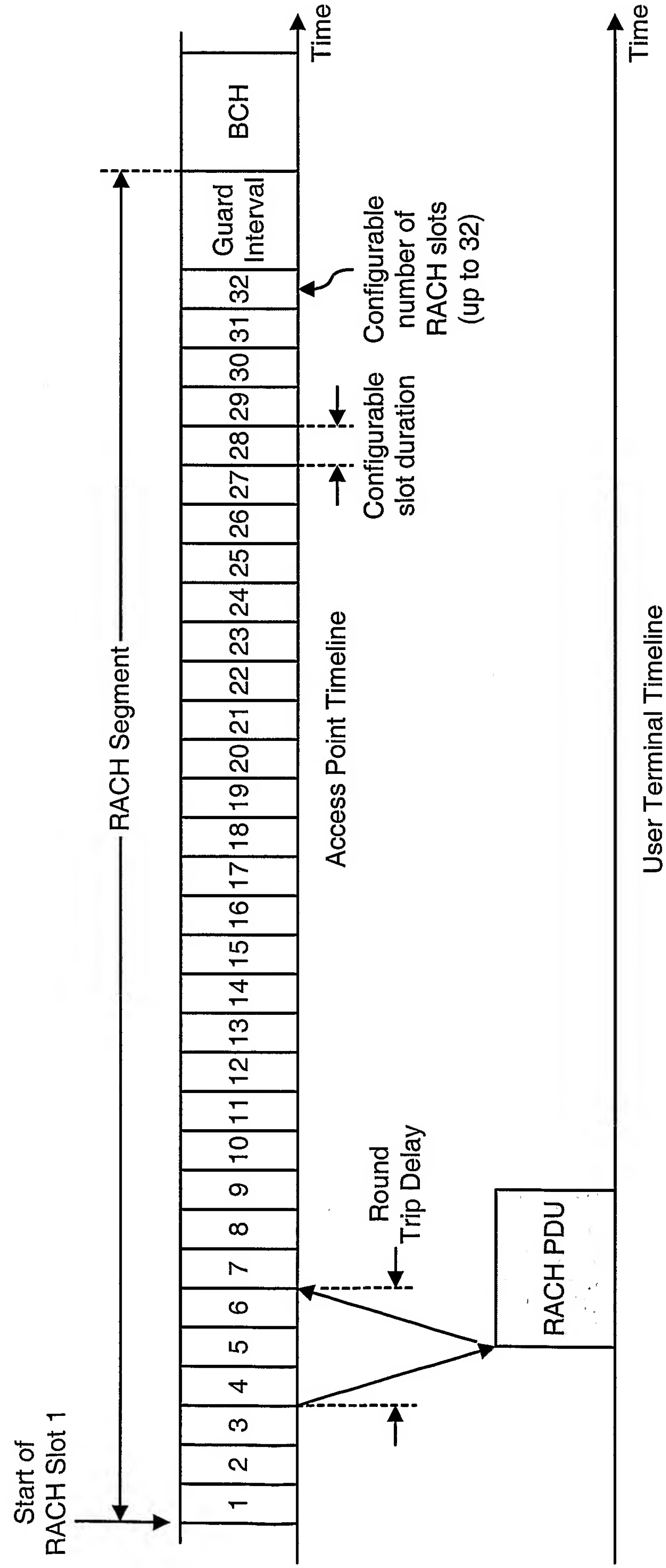

Start of
RACH Slot 1
RACH Segment
BCH
Guard
Interval
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20 21 22 23 24 25 26 27 28 29 30 31 32
Time
Configurable
number of
RACH slots
(up to 32)
Configurable
slot duration
Access Point Timeline
Round
Trip Delay
RACH PDU
Time
User Terminal Timeline

FIG. 13

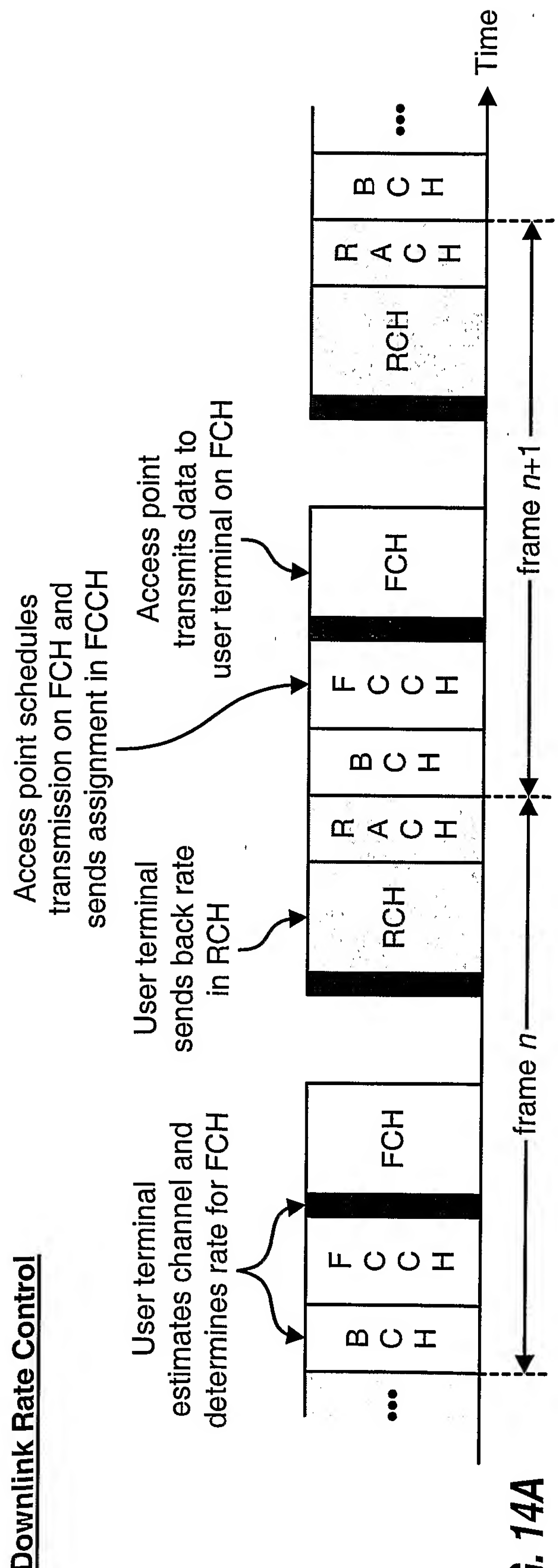


FIG. 14A

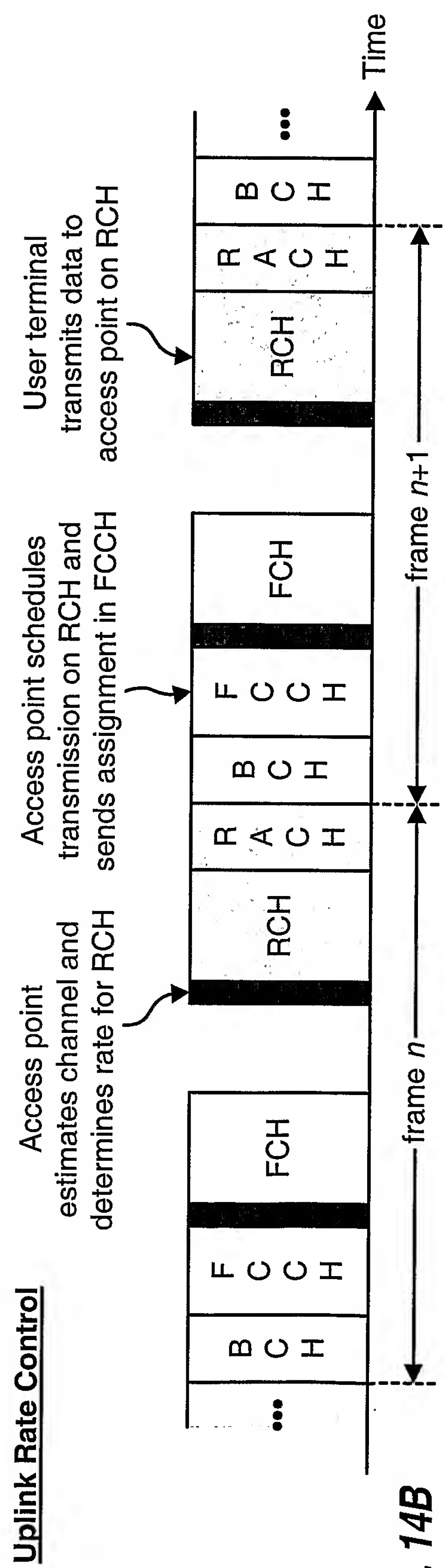


FIG. 14B

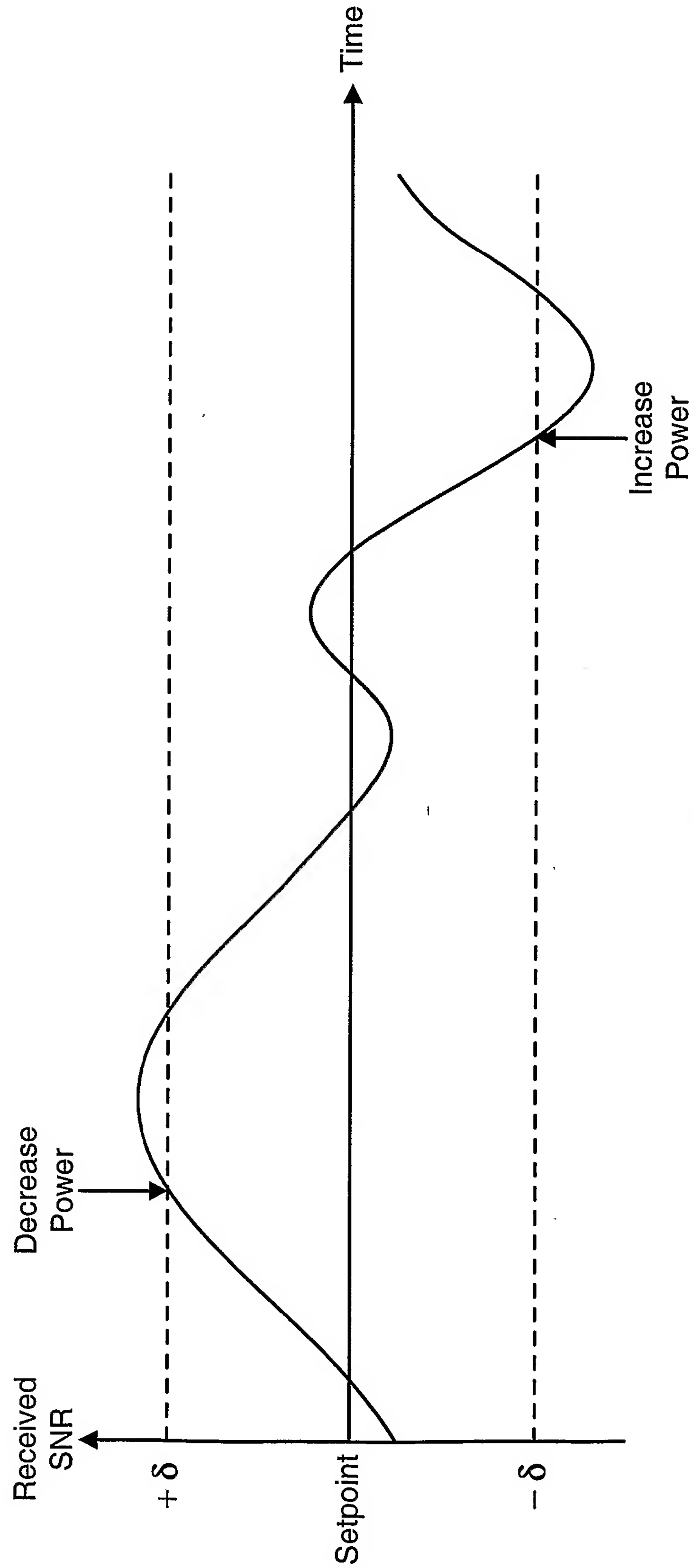

Time
Increase Power
Decrease Power
Received SNR
+ δ
Setpoint
– δ

FIG. 15

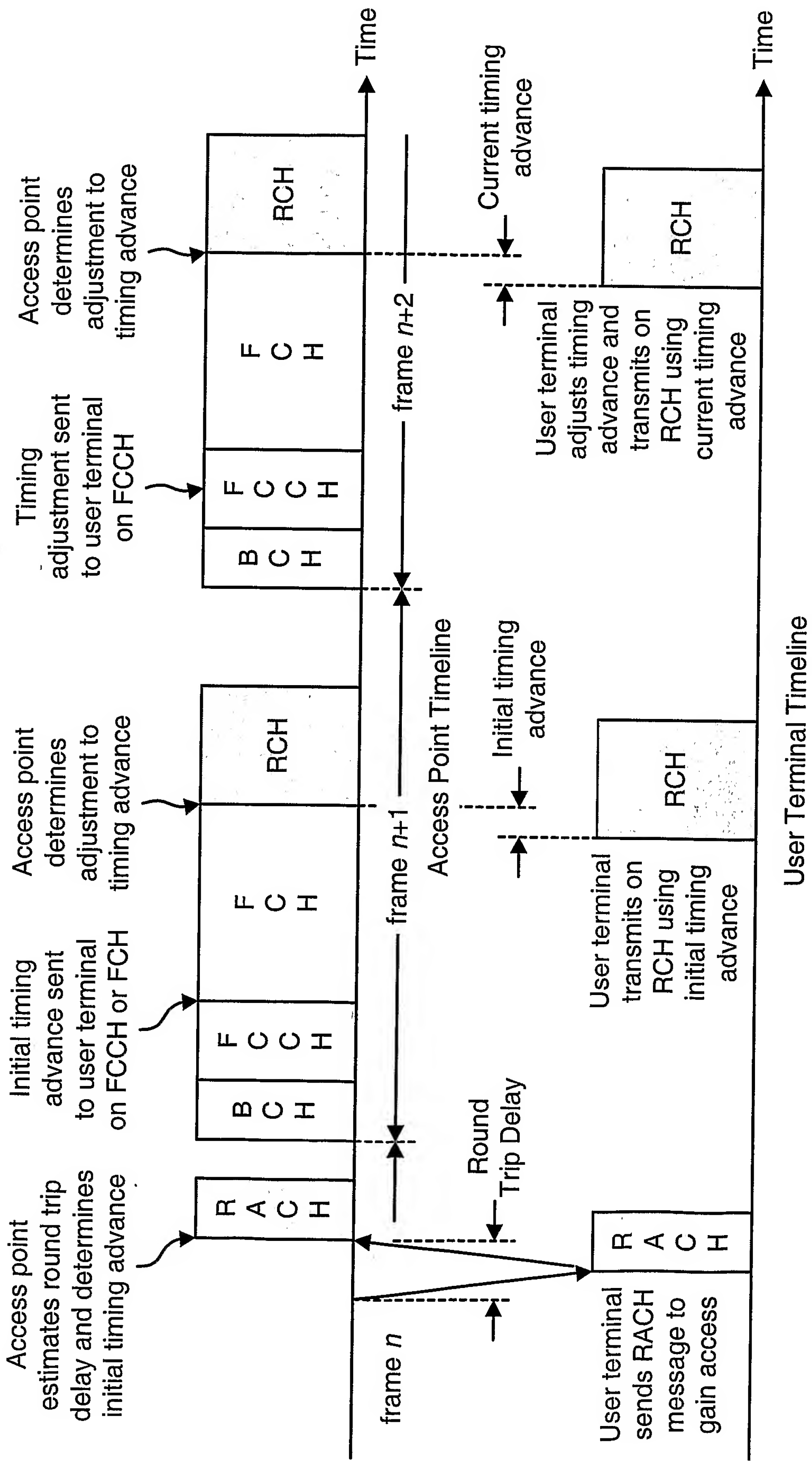

Access point estimates round trip delay and determines initial timing advance
Initial timing advance sent to user terminal on FCCH or FCH
Access point determines adjustment to timing advance
Timing adjustment sent to user terminal on FCCH
Access point determines adjustment to timing advance
RACH
BCCH
FCCH
FCH
RCH
BCCH
FCCH
FCH
RCH
Time
frame n
frame n+1
frame n+2
Access Point Timeline
Round Trip Delay
Initial timing advance
Current timing advance
RACH
RCH
RCH
Time
User terminal sends RACH message to gain access
User terminal transmits on RCH using initial timing advance
User terminal adjusts timing advance and transmits on RCH using current timing advance
User Terminal Timeline

FIG. 16

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(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, UZ, VC, VN, YU, ZA, ZM, ZW.

(84) Designated States *(regional)*: ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IT, LU, MC, NL, PT, RO, SE, SI, SK, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

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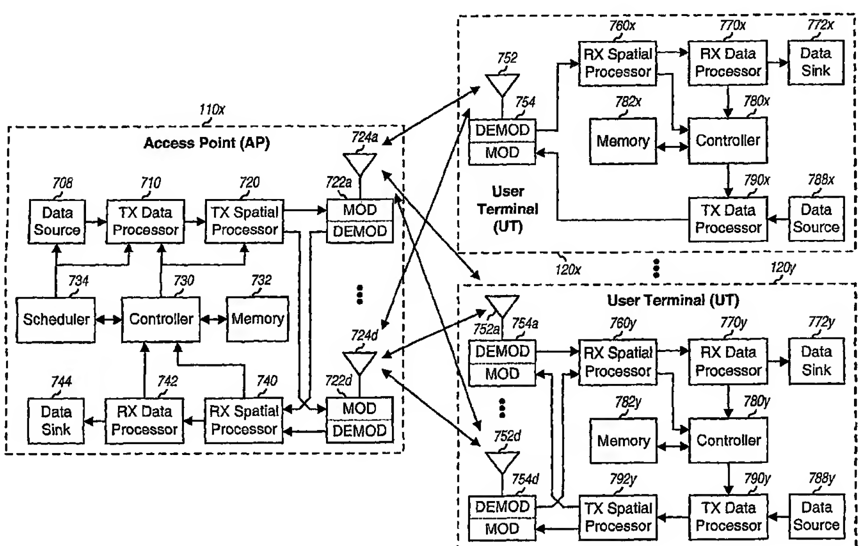
[Continued on next page]

(54) Title: MIMO WLAN SYSTEM



(57) Abstract: A multiple-access MIMO WLAN system that employs MIMO, OFDM, and TDD. The system (1) uses a channel structure with a number of configurable transport channels, (2) supports multiple rates and transmission modes, which are configurable based on channel conditions and user terminal capabilities, (3) employs a pilot structure with several types of pilot (e.g., beacon, MIMO, steered reference, and carrier pilots) for different functions, (4) implements rate, timing, and power control loops for proper system operation, and (5) employs random access for system access by the user terminals, fast acknowledgment, and quick resource assignments. Calibration may be performed to account for differences in the frequency responses of transmit/receive chains at the access point and user terminals. The spatial processing may then be simplified by taking advantage of the reciprocal nature of the downlink and uplink and the calibration.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 03/34514

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L12/28

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04B H04L H04Q H01Q H04J

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, INSPEC, COMPENDEX, IBM-TDB, WPI Data, PAJ

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A		126-143

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Y	US 5 982 327 A (VOOK ET AL) 9 November 1999 (1999-11-09) abstract column 8, line 3 – column 9, line 2; claim 9	59-102

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[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
21 March 2005	12. 05. 05

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Falò, L

INTERNATIONAL SEARCH REPORT

International Application No
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X	"Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; Data Link Control (DLC) Layer; Part 1: Basic Data Transport Functions; ETSI TS 101 761-1" ETSI STANDARDS, EUROPEAN TELECOMMUNICATIONS STANDARDS INSTITUTE, SOPHIA-ANTIPO, FR, vol. BR, no. V131, December 2001 (2001-12), XP014006627 ISSN: 0000-0001	103-125
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International Application No
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C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
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International application No.
PCT/US 03/34514

INTERNATIONAL SEARCH REPORT

Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This International Search Report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. ☐ Claims Nos.:
because they relate to parts of the International Application that do not comply with the prescribed requirements to such an extent that no meaningful International Search can be carried out, specifically:

3. ☐ Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

see additional sheet

1. ☒ As all required additional search fees were timely paid by the applicant, this International Search Report covers all searchable claims.

2. ☐ As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. ☐ As only some of the required additional search fees were timely paid by the applicant, this International Search Report covers only those claims for which fees were paid, specifically claims Nos.:

4. ☐ No required additional search fees were timely paid by the applicant. Consequently, this International Search Report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

☐ The additional search fees were accompanied by the applicant's protest.

☐ No prctest accompanied the payment of additional search fees.

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

This International Searching Authority found multiple (groups of) inventions in this international application, as follows:

1. claims: 1-40

   Method and apparatus for selecting transmission modes.

   ---

2. claims: 41-58

   Method and apparatus for time duplexing uplink and downlink channels.

   ---

3. claims: 59-102

   Method and apparatus for performing spatial processing on a data transmission.

   ---

4. claims: 103-125

   Channel structure for wireless data transmission and apparatus thereof.

   ---

5. claims: 126-143

   Method and apparatus for transmitting and receiving signalling information.

   ---

6. claims: 144-166

   Method and apparatus for coding data frames.

   ---

7. claims: 167-187

   Method and apparatus for network access.

   ---

8. claims: 188-216

   Method and apparatus for determining data rate based on channel estimation.

INTERNATIONAL SEARCH REPORT

Information on patent family members

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Information on patent family members

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